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Spahlinger et al.

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(54) **CONTROLLER UNIT AND DEVICE FOR RESETTING AN OSCILLATOR EXCITED BY A HARMONIC OSCILLATION, AND YAW RATE SENSOR**

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(73) Assignee: **Northrop Grumman Litef GmbH** (DE)

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H03L 7/00 (2006.01)
G05B 5/01 (2006.01)

(52) **U.S. Cl.**
CPC ... ***H03L 7/00*** (2013.01); ***G05B 5/01*** (2013.01)

(58) **Field of Classification Search**
CPC H03L 7/00; G05D 19/02; G01C 19/00
USPC 331/25; 73/504.01, 504.02, 504.03
See application file for complete search history.

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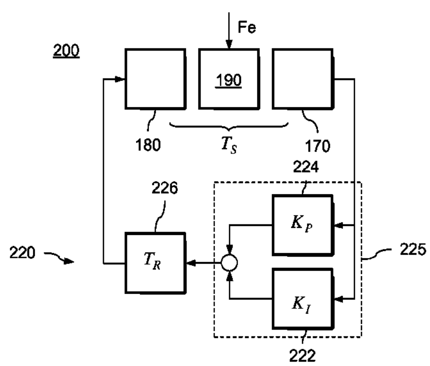
(74) *Attorney, Agent, or Firm* — Tarolli, Sundheim, Covell & Tummino LLP

(57) **ABSTRACT**

A controller unit includes a PI-controller for harmonic command variables. The transfer function of the PI-controller has a conjugate complex pole at a controller angular frequency ω_r in the s-plane or a pole at $e^{\pm j\omega_r T}$ in the z-plane, wherein T is the sampling time of a discrete input signal of the PI-controller and ω_r is larger than 0. The controller angular frequency ω_r is chosen equal to the resonance angular frequency ω_0 of an oscillator. The controller parameters, are, for example determined by pole/zero compensation. The controller unit allows, for example, a broad band control of harmonic oscillators in rotation rate sensors.

18 Claims, 18 Drawing Sheets



$$\frac{K_I}{K_P} \approx s_0 \quad vz = sign(K_I)$$

$$(T_S + T_R)\cdot\omega_0 = \frac{3}{2}\pi \quad \textit{für} \quad vz = +1$$

$$(T_S + T_R)\cdot\omega_0 = \frac{1}{2}\pi \quad \textit{für} \quad vz = -1$$

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Fig. 1A
Prior Art

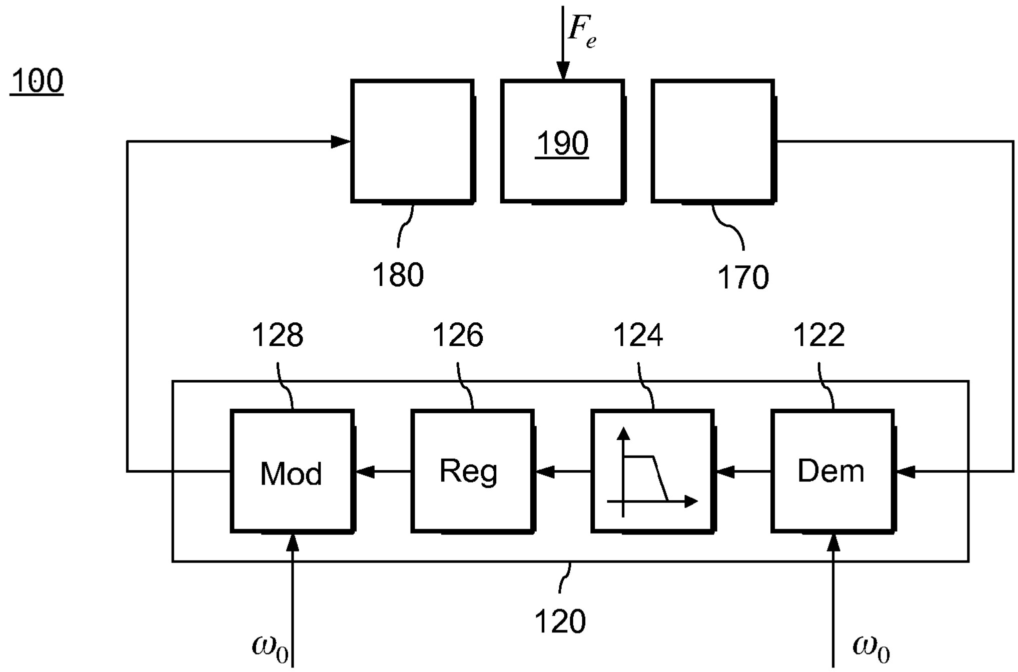


Fig. 1B
Prior Art

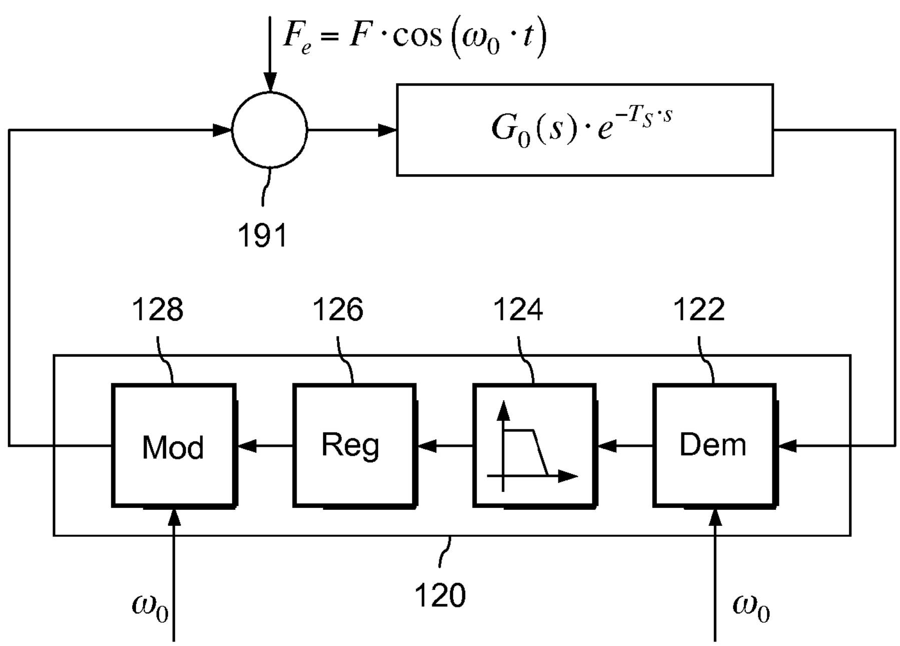


Fig. 1C
Prior Art

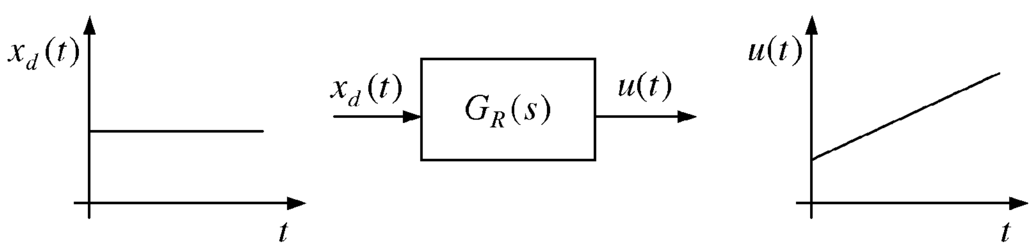
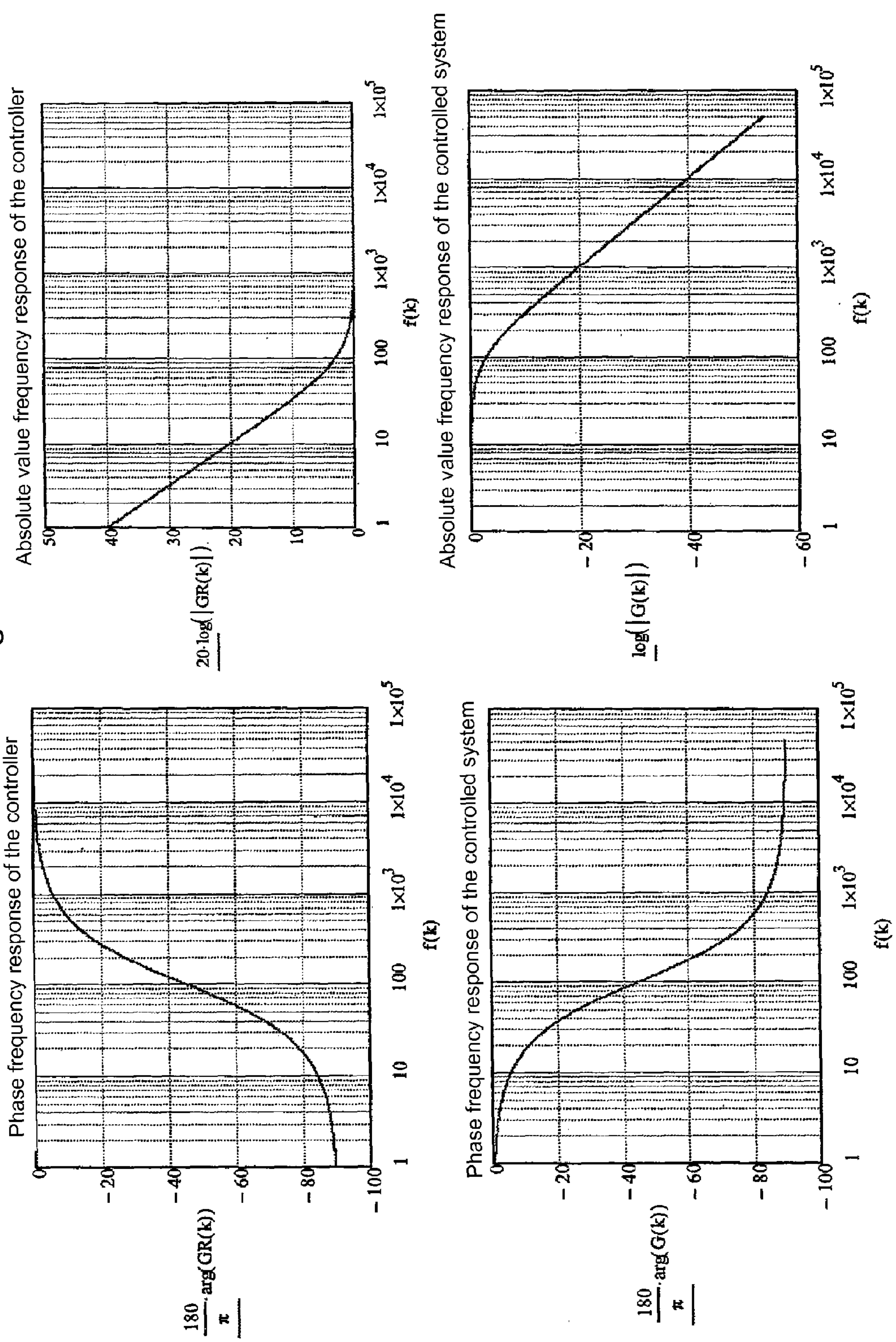


Fig. 1D


Absolute value frequency response of the controller
20·log(|GR(k)|)
f(k)
Absolute value frequency response of the controlled system
log(|G(k)|)
f(k)
Phase frequency response of the controller
180/π · arg(GR(k))
f(k)
Phase frequency response of the controlled system
180/π · arg(G(k))
f(k)

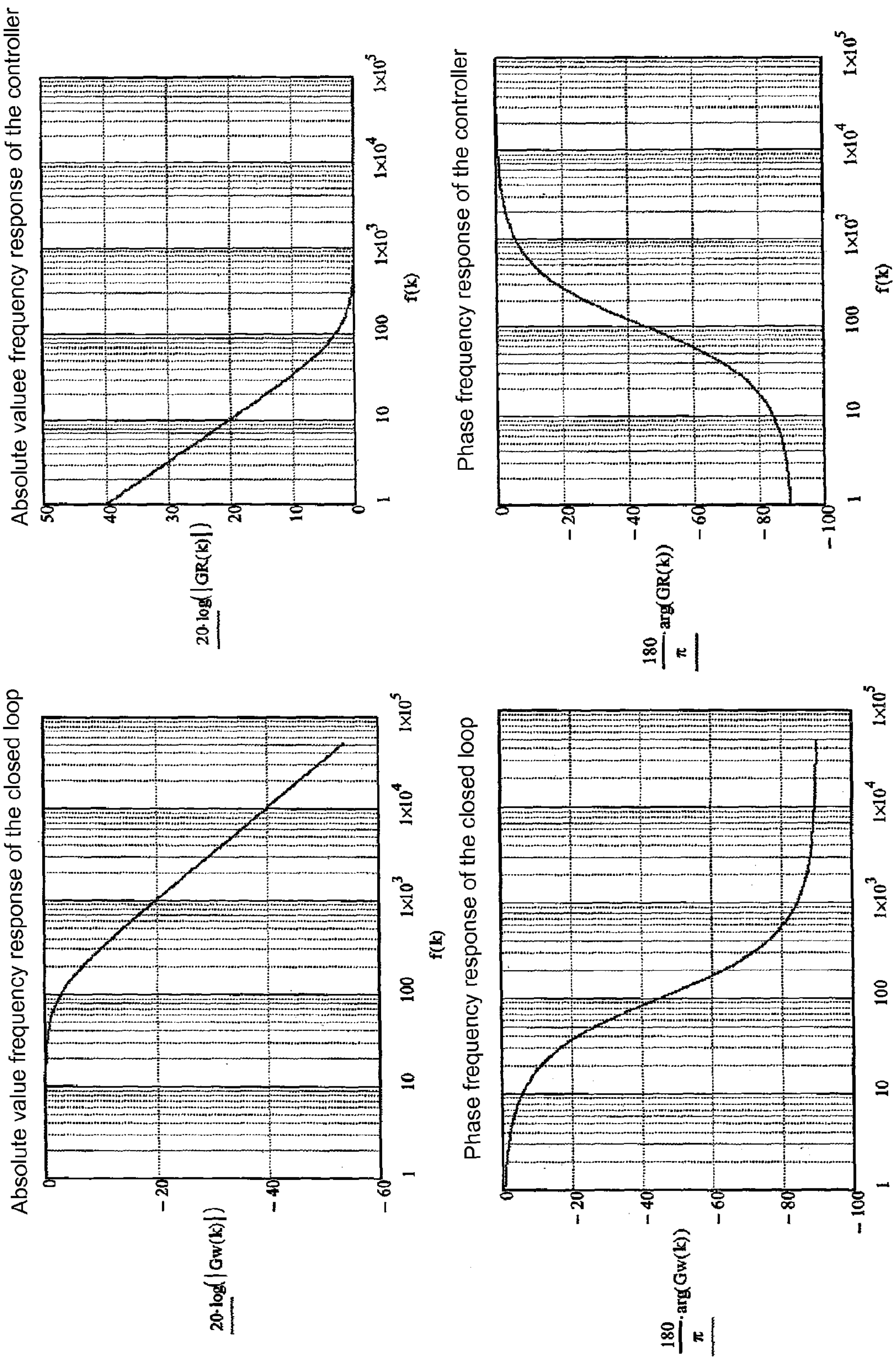
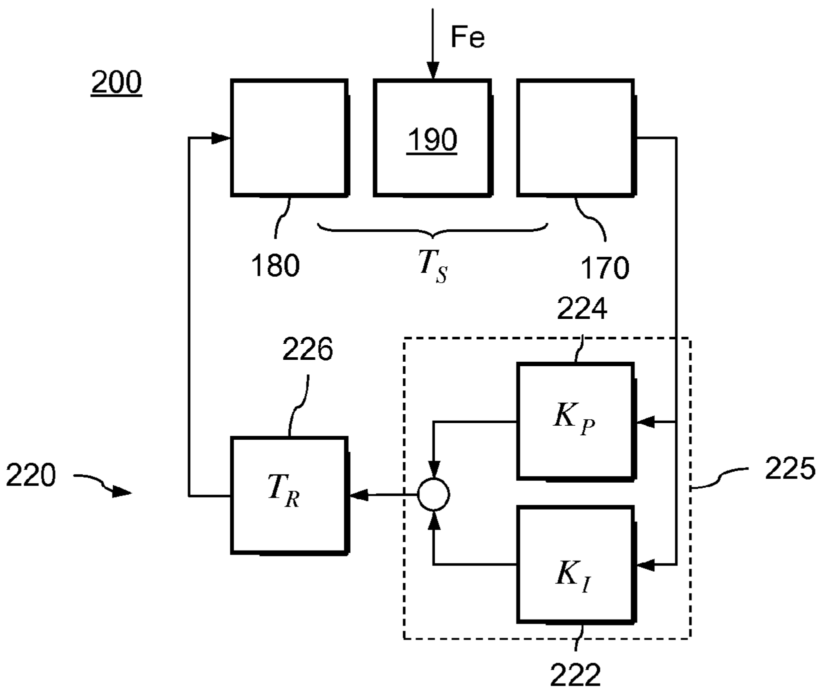

Absolute valuee frequency response of the controller
$20 \cdot \log(|GR(k)|)$
f(k)
Phase frequency response of the controller
$\frac{180}{\pi} \cdot \arg(GR(k))$
f(k)
Absolute value frequency response of the closed loop
$20 \cdot \log(|Gw(k)|)$
f(k)
Phase frequency response of the closed loop
$\frac{180}{\pi} \cdot \arg(Gw(k))$
f(k)

Fig. 1D (Cont)

Fig. 2A



$$\frac{K_I}{K_P} \approx s_0 \qquad vz = sign(K_I)$$

$$(T_S + T_R) \cdot \omega_0 = \frac{3}{2}\pi \quad \text{für} \quad vz = +1$$

$$(T_S + T_R) \cdot \omega_0 = \frac{1}{2}\pi \quad \text{für} \quad vz = -1$$

Fig. 2B

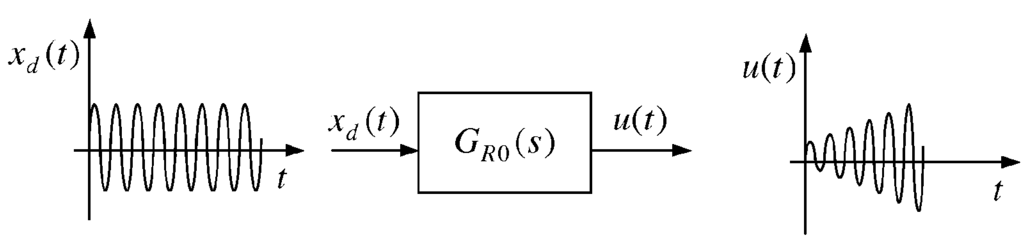
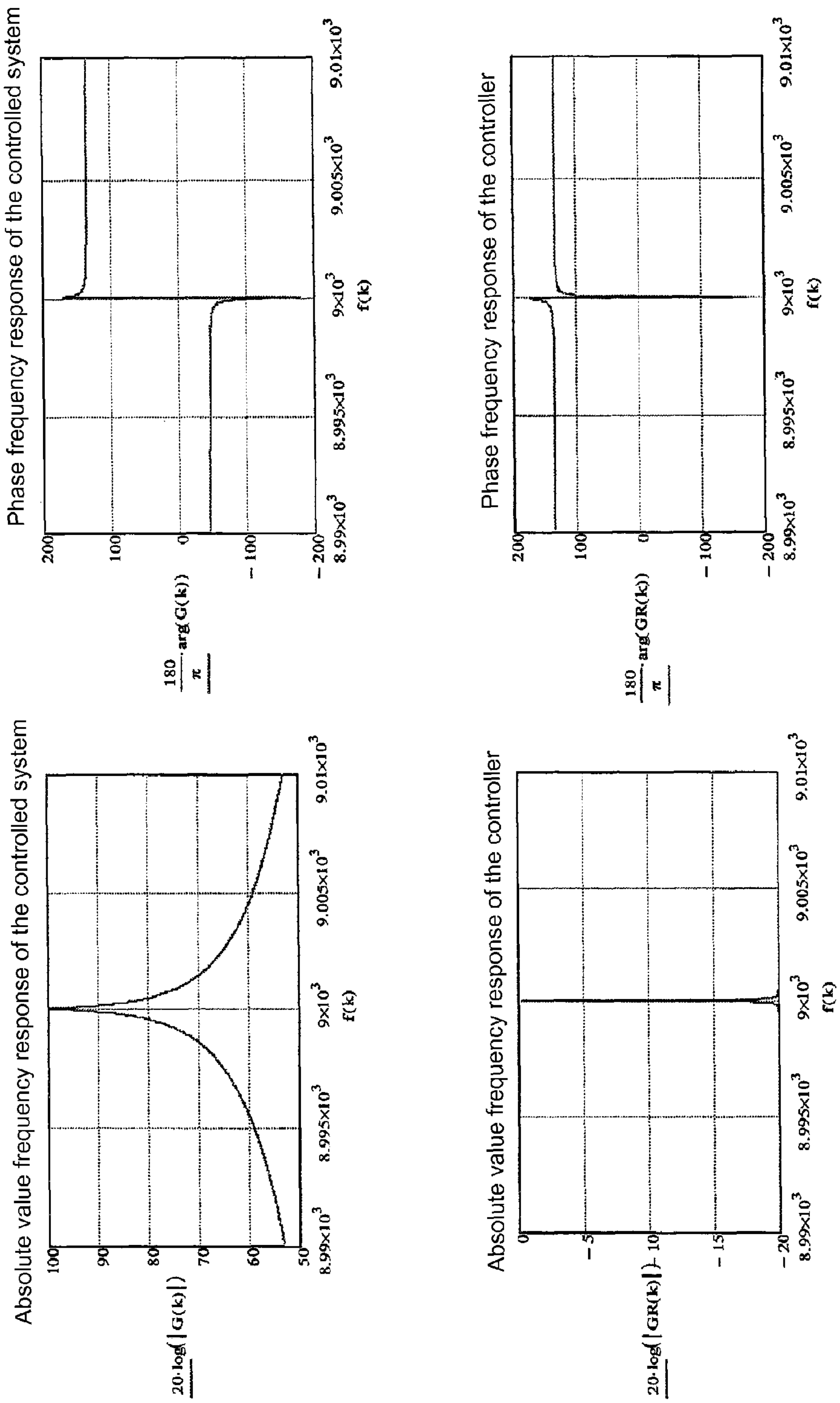


Fig. 2C



Phase frequency response of the controlled system
Phase frequency response of the controller
Absolute value frequency response of the controlled system
Absolute value frequency response of the controller
$\frac{180}{\pi} \cdot \arg(G(k))$
$\frac{180}{\pi} \cdot \arg(GR(k))$
$20 \cdot \log(|G(k)|)$
$20 \cdot \log(|GR(k)|) - 10$
f(k)
8.99×10^3
8.995×10^3
9×10^3
9.005×10^3
9.01×10^3
200
100
0
−100
−200
100
90
80
70
60
50
0
−5
−10
−15
−20

Fig. 2C (Cont.)

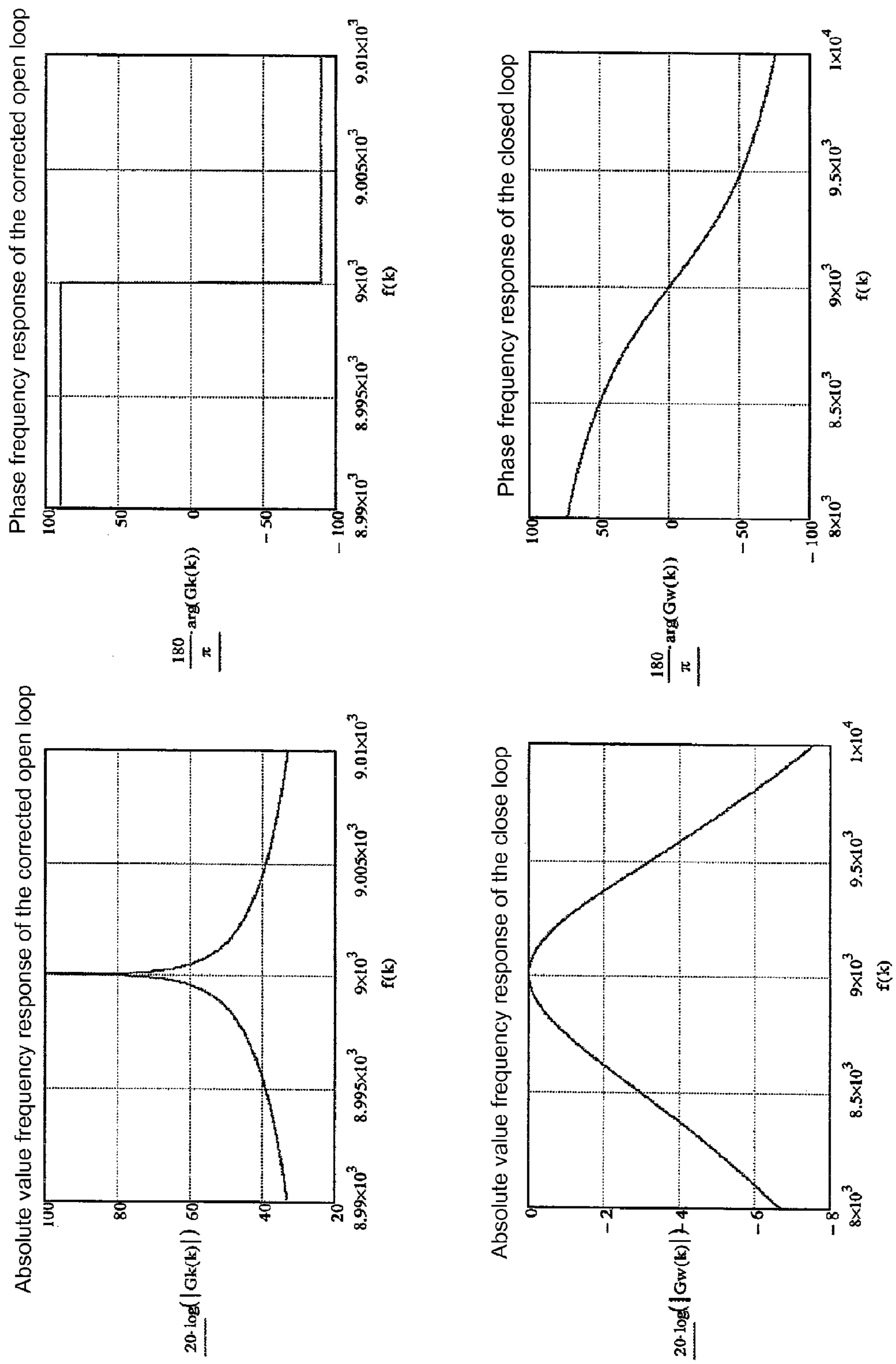
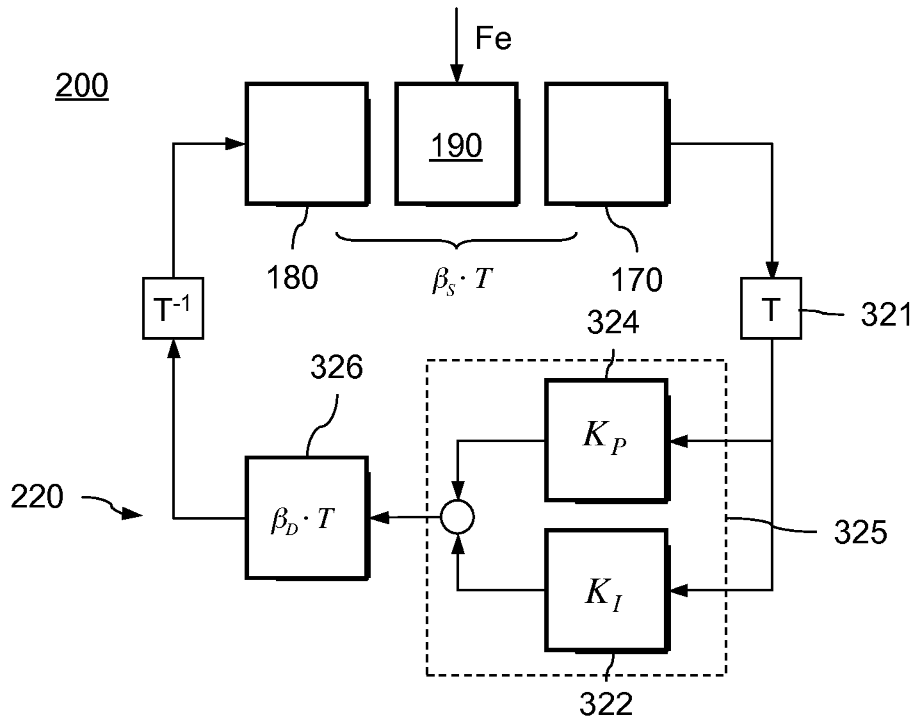

Absolute value frequency response of the corrected open loop
20·log(|Gk(k)|)
100
80
60
40
20
8.99×10^3
8.995×10^3
9×10^3
9.005×10^3
9.01×10^3
f(k)
Phase frequency response of the corrected open loop
180/π·arg(Gk(k))
100
50
0
−50
−100
8.99×10^3
8.995×10^3
9×10^3
9.005×10^3
9.01×10^3
f(k)
Absolute value frequency response of the close loop
20·log(|Gw(k)|)
0
−2
−4
−6
−8
8×10^3
8.5×10^3
9×10^3
9.5×10^3
1×10^4
f(k)
Phase frequency response of the closed loop
180/π·arg(Gw(k))
100
50
0
−50
−100
8×10^3
8.5×10^3
9×10^3
9.5×10^3
1×10^4
f(k)

Fig. 3A

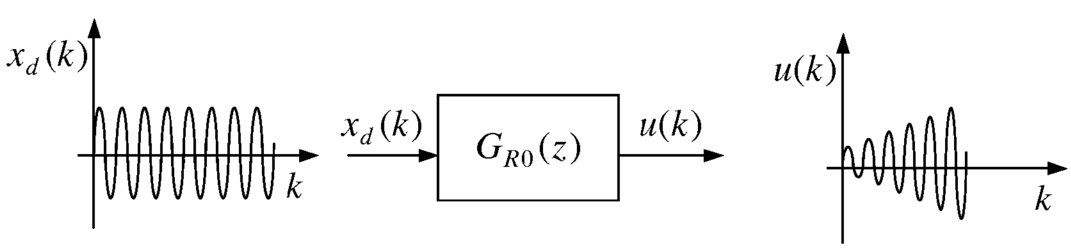


$$\frac{K_I}{K_P} \approx s_0 \quad vz = sign(K_I)$$

$$\left(\beta_S + \beta_D + \frac{1}{2}\right) \cdot \omega_0 \cdot T = \frac{3}{2}\pi \quad für \quad vz = +1$$

$$\left(\beta_S + \beta_D + \frac{1}{2}\right) \cdot \omega_0 \cdot T = \frac{1}{2}\pi \quad für \quad vz = -1$$

Fig. 3B



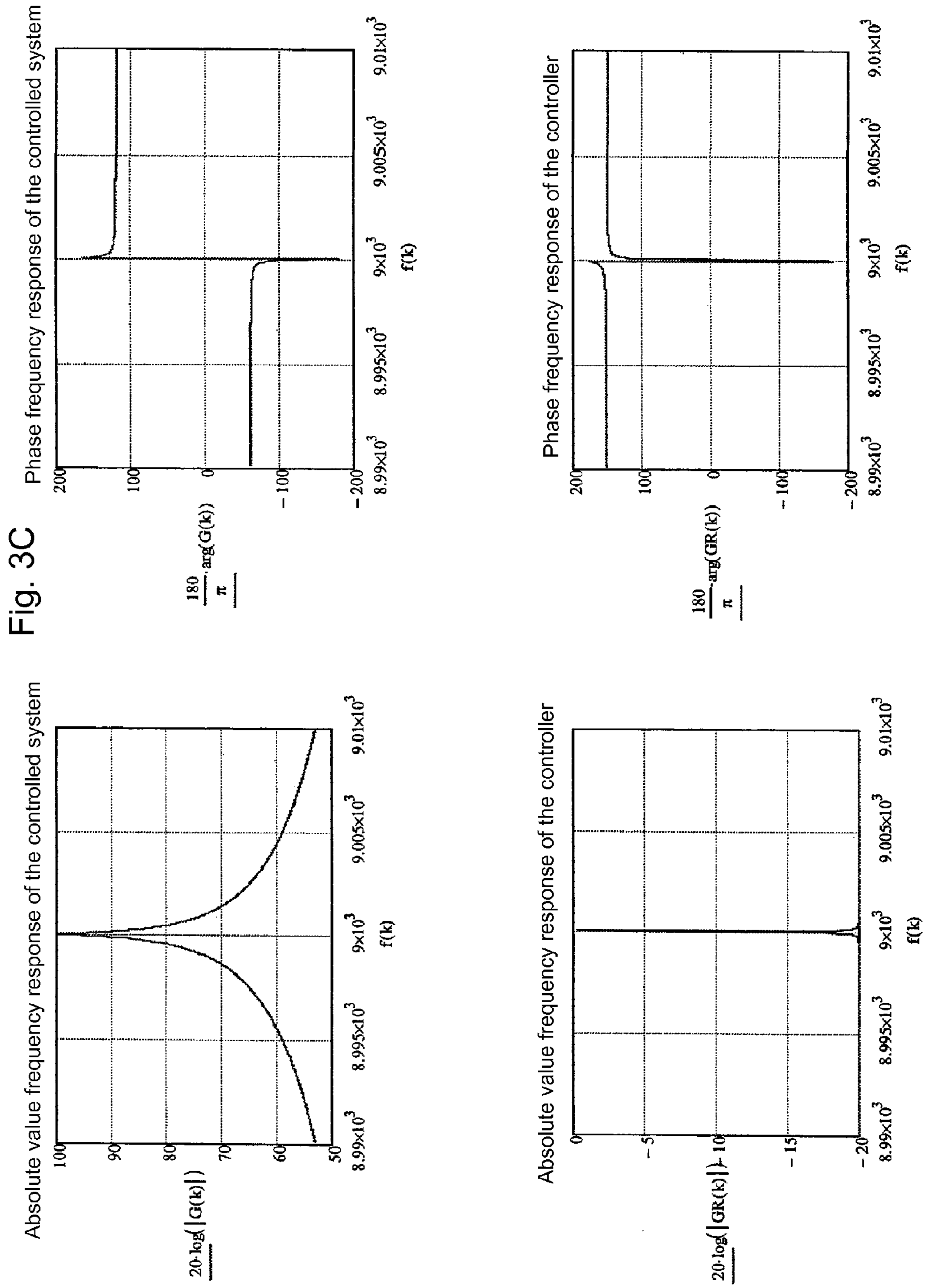

Phase frequency response of the controlled system
Phase frequency response of the controller
Absolute value frequency response of the controlled system
Absolute value frequency response of the controller
$\frac{180}{\pi}\cdot\arg(G(k))$
$\frac{180}{\pi}\cdot\arg(GR(k))$
$20\cdot\log(|G(k)|)$
$20\cdot\log(|GR(k)|)-10$
f(k)
8.99×10^3
8.995×10^3
9×10^3
9.005×10^3
9.01×10^3

Fig. 3C

Fig. 3C (Cont.)

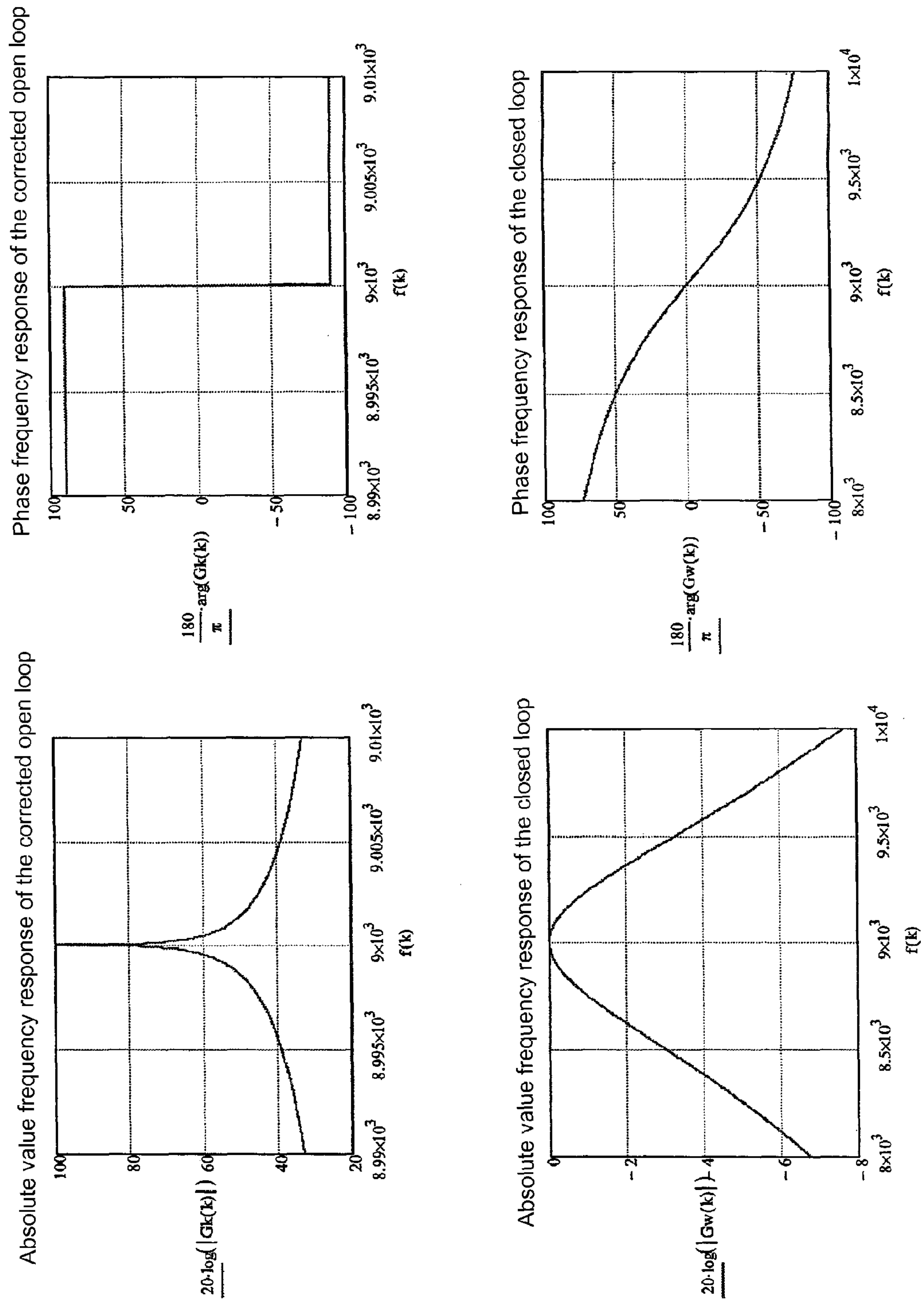
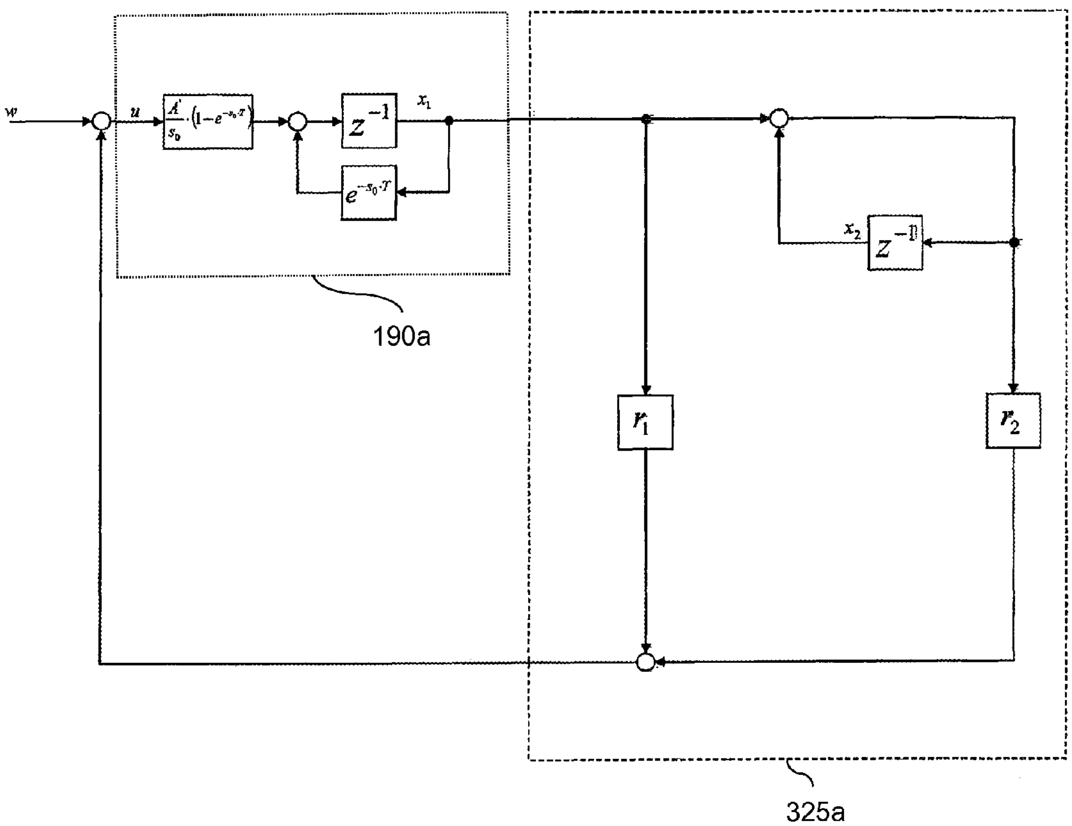

Absolute value frequency response of the corrected open loop
20·log(|Gk(k)|)
f(k)
Phase frequency response of the corrected open loop
180/π·arg(Gk(k))
f(k)
Absolute value frequency response of the closed loop
20·log(|Gw(k)|)
f(k)
Phase frequency response of the closed loop
180/π·arg(Gw(k))
f(k)

Fig. 3D



w
u
$\frac{A}{s_0}\cdot(1-e^{-s_0 \cdot T})$
z^{-1}
x_1
$e^{-s_0 \cdot T}$
x_2
z^{-1}
r_1
r_2
190a
325a

Fig. 3E

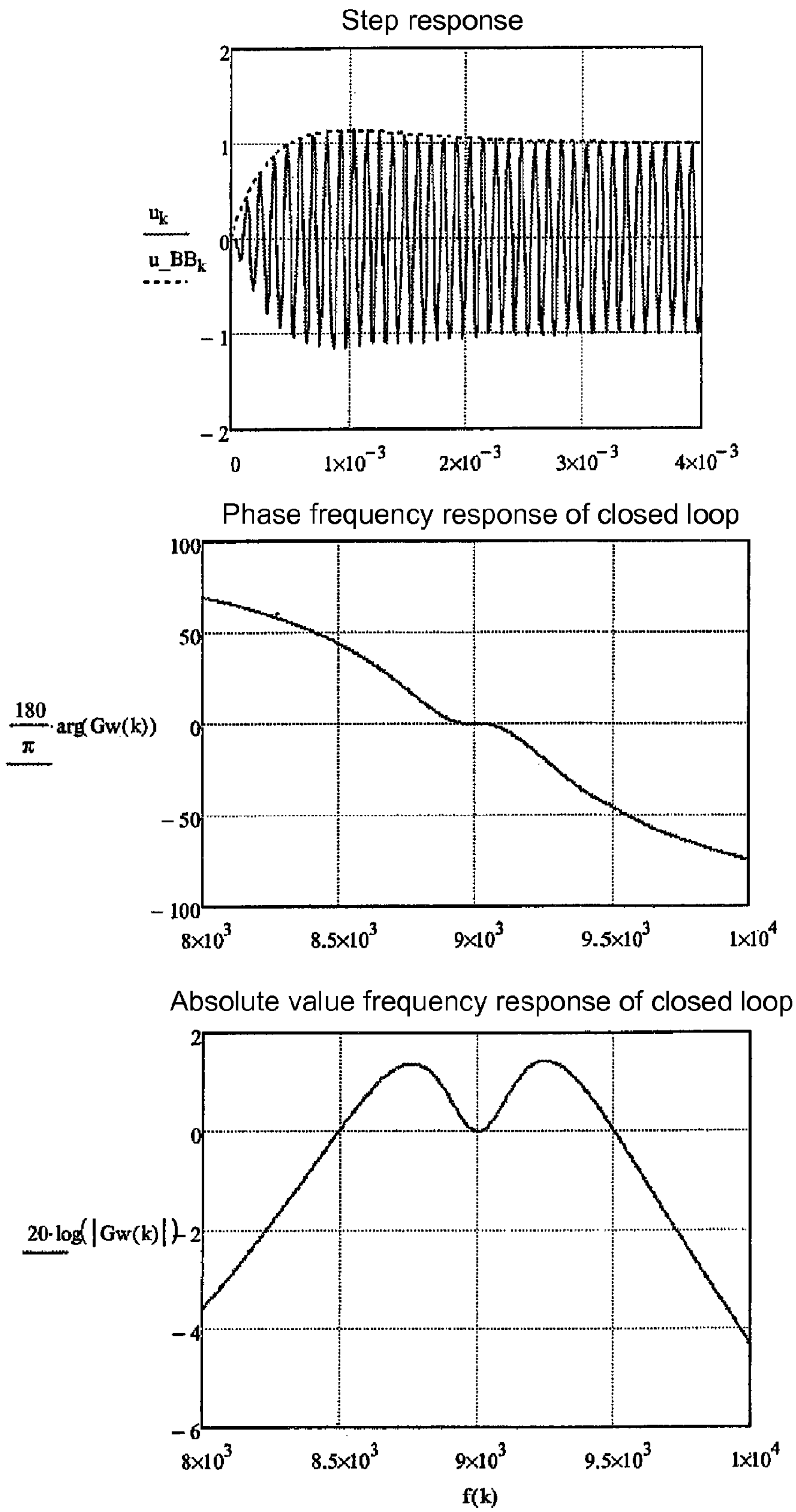
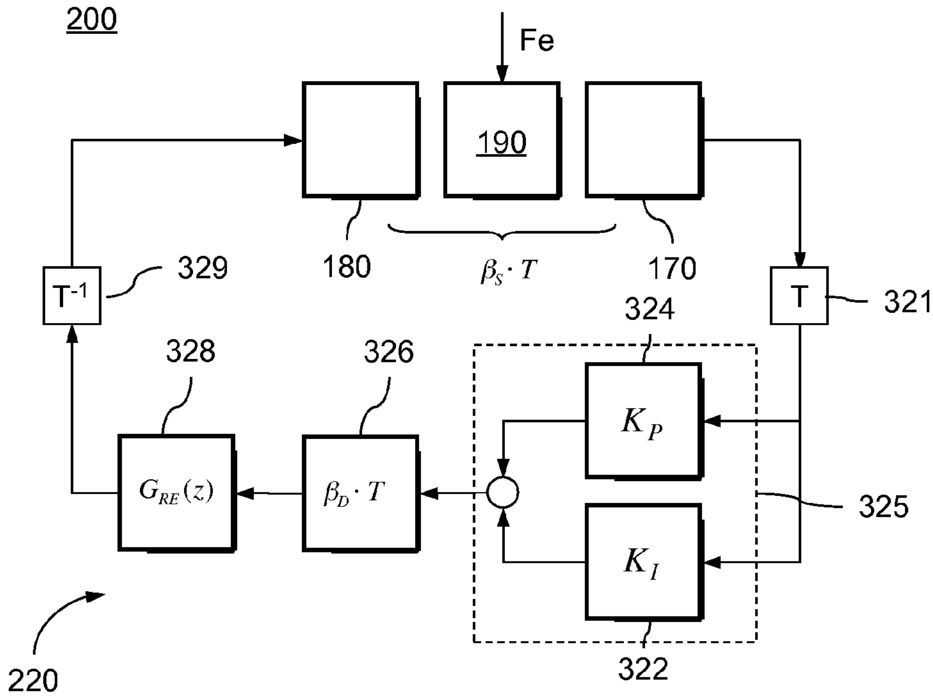
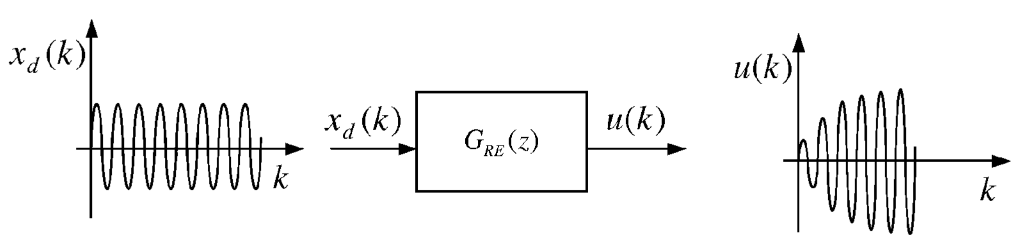

Step response
2
1
0
- 1
- 2
u_k
u_BB_k
0
1×10^{-3}
2×10^{-3}
3×10^{-3}
4×10^{-3}
Phase frequency response of closed loop
100
50
0
- 50
- 100
$\frac{180}{\pi}\cdot arg(Gw(k))$
8×10^{3}
8.5×10^{3}
9×10^{3}
9.5×10^{3}
1×10^{4}
Absolute value frequency response of closed loop
2
0
- 2
- 4
- 6
$20\cdot\log(|Gw(k)|)$
8×10^{3}
8.5×10^{3}
9×10^{3}
9.5×10^{3}
1×10^{4}
f(k)

Fig. 4A



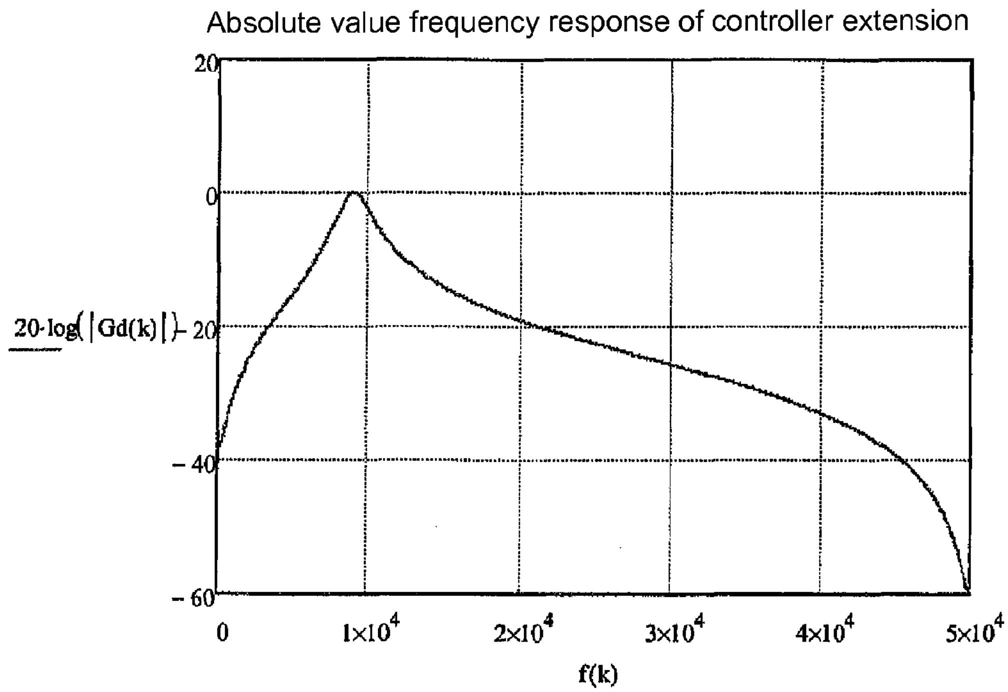
200
Fe
190
180
$\beta_S \cdot T$
170
329
T^{-1}
T
321
324
328
326
K_P
$G_{RE}(z)$
$\beta_D \cdot T$
325
K_I
220
322

Fig. 4B



$x_d(k)$
k
$x_d(k)$
$G_{RE}(z)$
$u(k)$
$u(k)$
k

Fig. 4C


Absolute value frequency response of controller extension
20
0
20·log(|Gd(k)|) − 20
− 40
− 60
0
1×10^4
2×10^4
3×10^4
4×10^4
5×10^4
f(k)

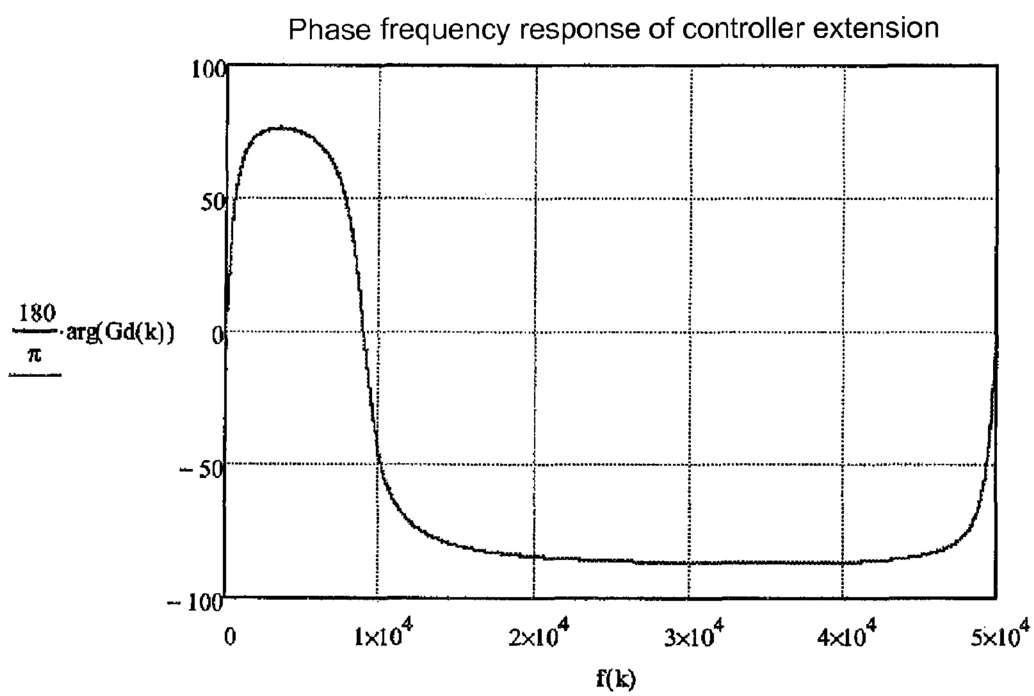

Phase frequency response of controller extension
100
50
180/π·arg(Gd(k)) 0
− 50
− 100
0
1×10^4
2×10^4
3×10^4
4×10^4
5×10^4
f(k)

Fig. 5A

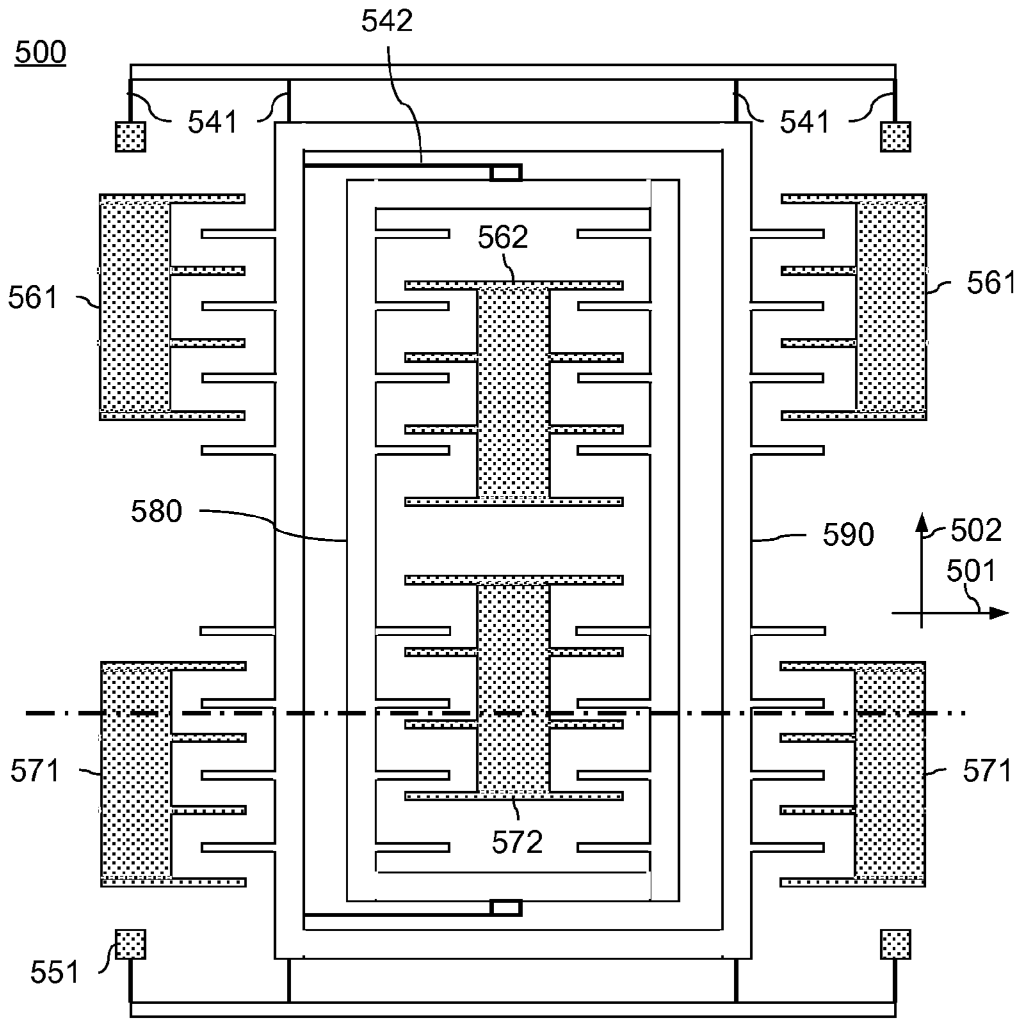

500
542
541
541
561
562
561
580
590
502
501
571
571
572
551

Fig. 5B

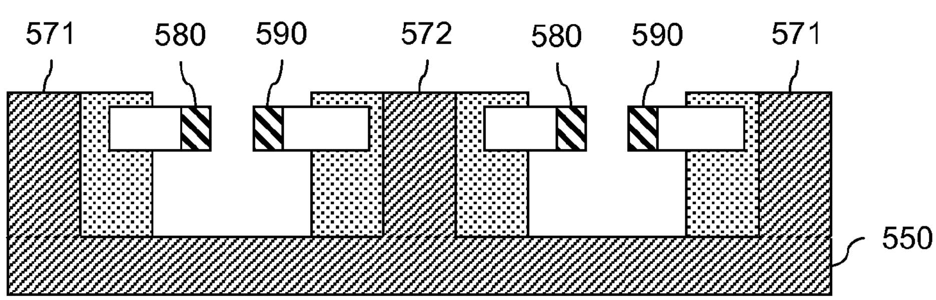

571
580
590
572
580
590
571
550

Fig. 5C

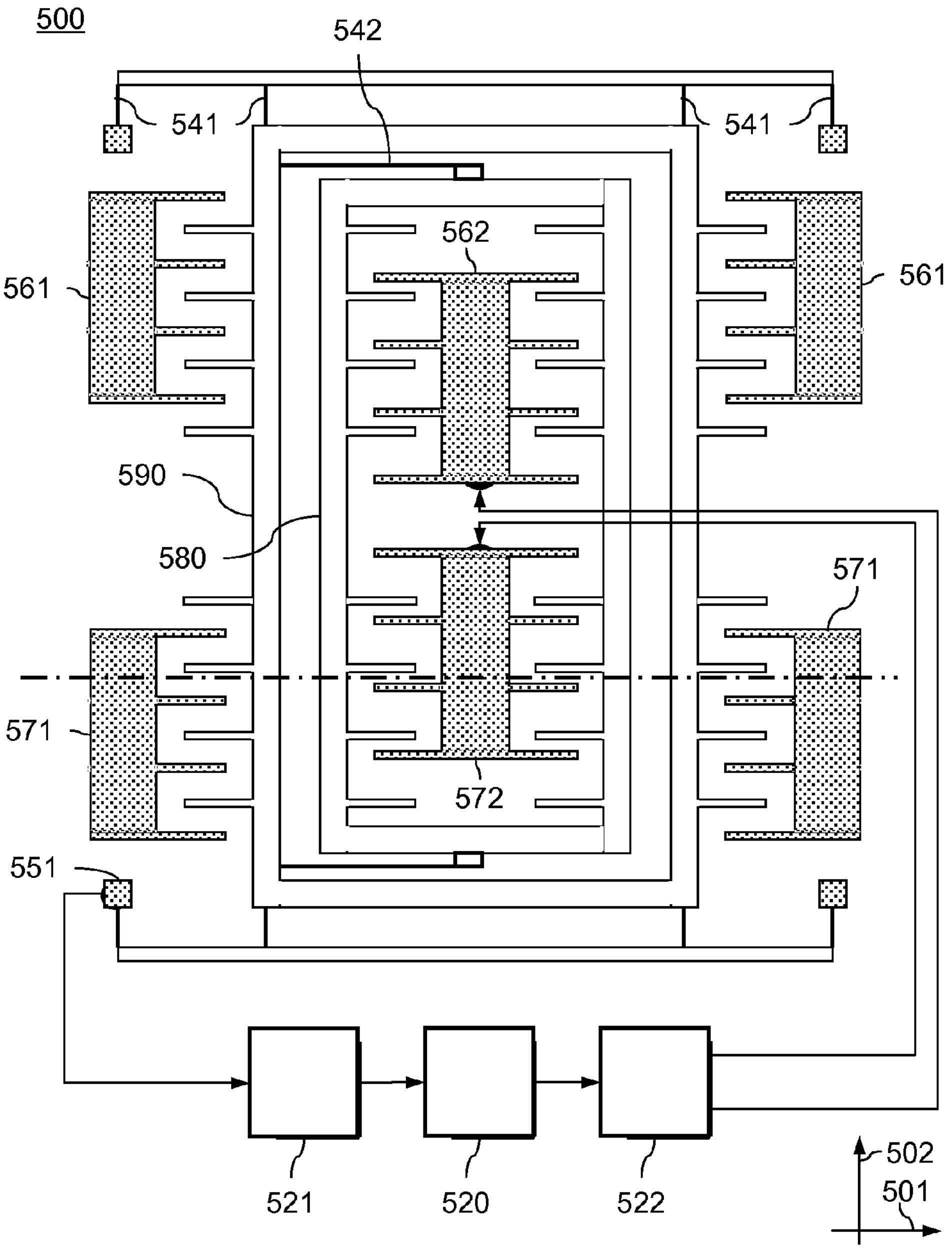

500
542
541
541
562
561
561
590
580
571
571
551
572
521
520
522
502
501

Fig. 6

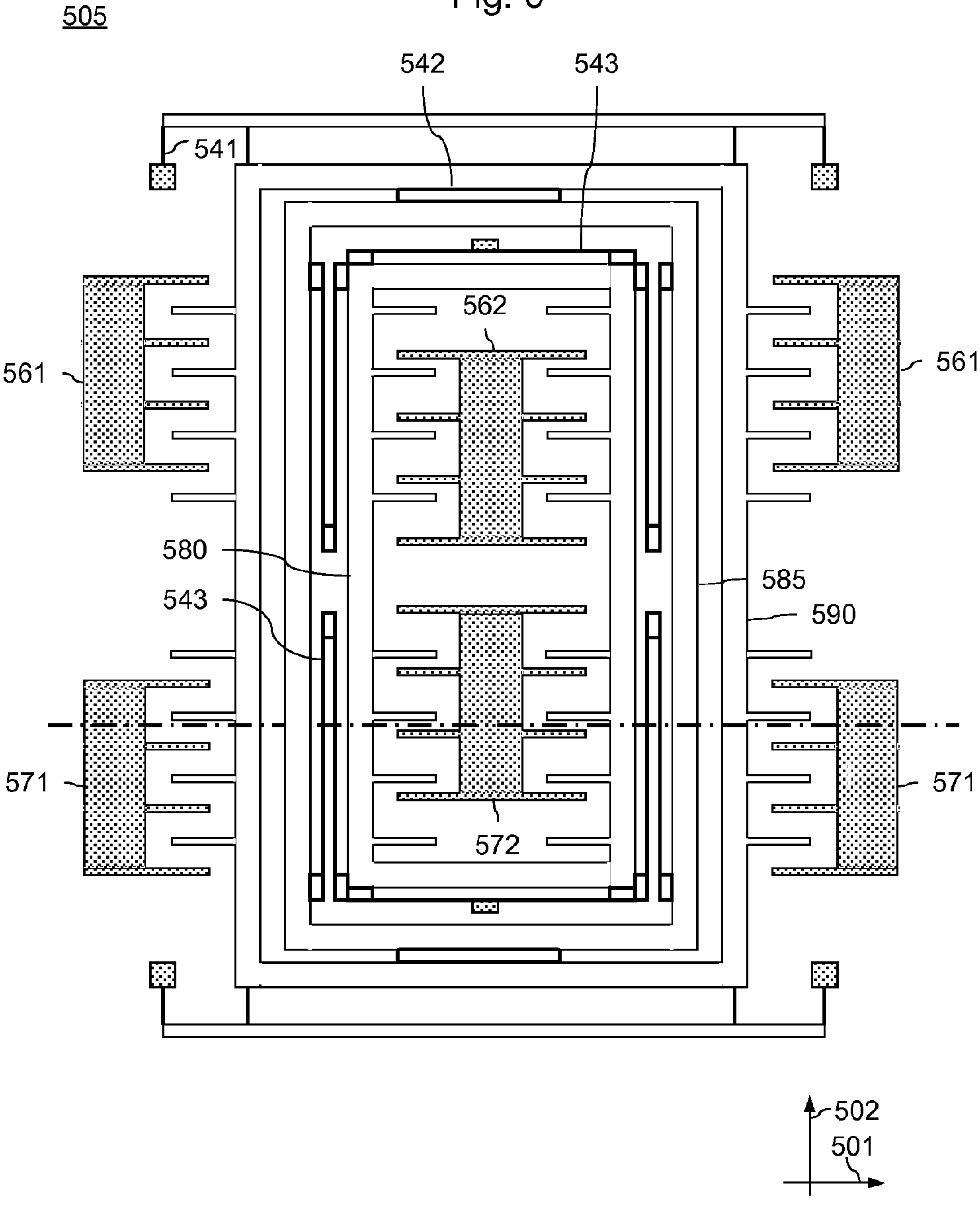


Fig. 7A

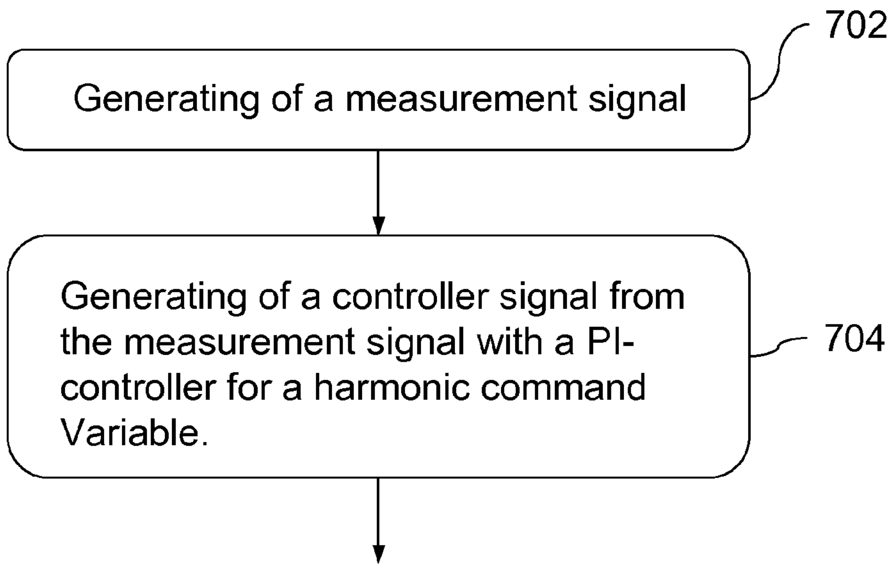
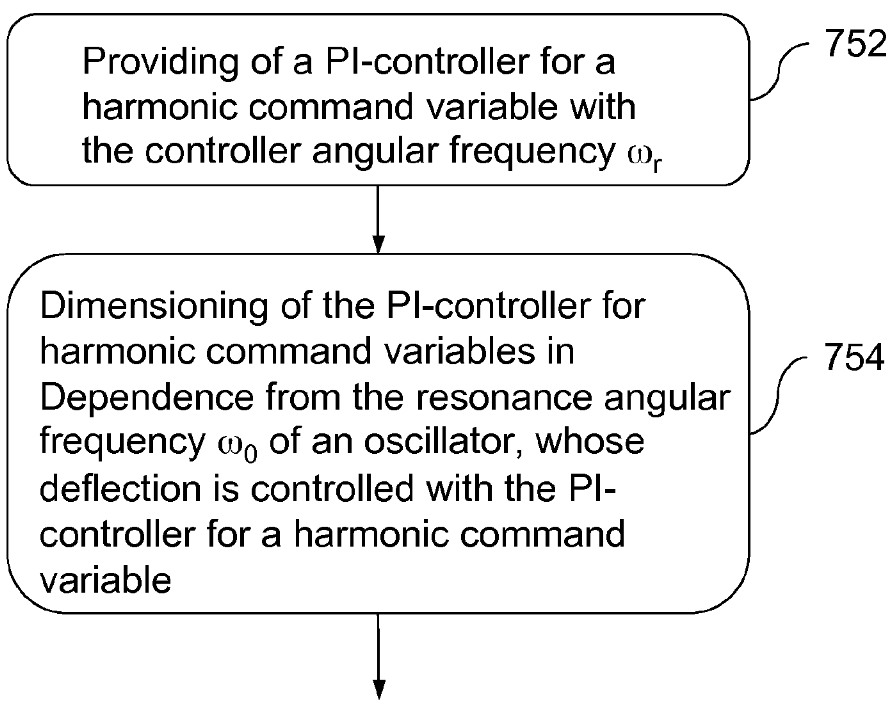

702
Generating of a measurement signal
Generating of a controller signal from the measurement signal with a PI-controller for a harmonic command Variable.
704

Fig. 7B


752
Providing of a PI-controller for a harmonic command variable with the controller angular frequency ω_r
Dimensioning of the PI-controller for harmonic command variables in Dependence from the resonance angular frequency ω_0 of an oscillator, whose deflection is controlled with the PI-controller for a harmonic command variable
754

CONTROLLER UNIT AND DEVICE FOR RESETTING AN OSCILLATOR EXCITED BY A HARMONIC OSCILLATION, AND YAW RATE SENSOR

BACKGROUND

1. Field of the Invention

The invention refers to a controller unit for resetting a deflection of an oscillator excited with a harmonic oscillation, a device including such a controller unit, in particular a rotation rate sensor, as well as to a method for operating and for manufacturing such a controller unit.

2. Description of the Prior Art

Conventional control methods are tailored to control problems with constant or only slowly changing command variables. The value of a controlled process variable affected from a disturbance is kept close to a predetermined set point, or is updated as close as possible to a changing set point. Some applications (e.g. micromechanical rotation rate sensors for analysis of a Coriolis force) provide a control loop for resetting a deflection of an oscillator oscillating with its resonance frequency when stationary. A controller for such a control loop with a harmonic oscillation as command variable is conventionally designed such that a harmonic force signal exciting the oscillator is compensated so that the oscillator—apart form the harmonic oscillation corresponding to the command variable—performs as little movement as possible.

Typically, this feedback control problem is solved as illustrated in FIGS. **1**A to **1**D. FIG. **1**A refers to a device **100** with a controlled system such as a mechanical oscillator **190**, whose translational or rotational deflection is captured by a sensor **170**. The oscillator **190** is supported or suspended such that it is movable along a direction of excitation and able to oscillate with a resonance angular frequency ω_0. A harmonic force signal F_e acts on the oscillator **190** along the direction of excitation. A measurement signal from the sensor **170** reproduces the movement of the oscillator **190** along the direction of excitation. The movement of the oscillator **190** includes a resonance oscillation with the resonance angular frequency ω_0, modulated by a force F (disturbance).

The measurement signal (system output signal) is fed to a controller unit **120** with a demodulator **122**. In the demodulator **122**, the system output signal is multiplied with a harmonic signal of frequency ω_0 which is equal to the resonance angular frequency ω_0 of the oscillator **190**, wherein a baseband version of the system output signal as well as additional frequency conversion products are formed. A low pass filter **124** damps higher frequency components, in particular at the double resonance angular frequency $2\cdot\omega_0$ of the oscillator **190**. The baseband signal is fed to the controller **126**, which operates in the baseband, whose design and dimensions can be established by known controller design methods. The controller **126** is, for example, a continuous PI-controller. Due to its integral component, high stationary position can be achieved in case of a constant command variable.

The output of the controller **125** is multiplied (modulated) with a harmonic signal of frequency ω_0 equal to the resonance angular frequency □$_0$ of the oscillator **190** in a modulator **128**. The modulation product is fed to an actuator **180** as a controller signal, the actuator executing according to the controller signal a force to the oscillator **190** that acts opposite to the deflection of the oscillator **190**. With the resonance angular frequency ω_0 and the damping s_0 of the oscillator as well as with the amplification A and the system dead time T_S of the system formed of the actuator **180**, the oscillator **190** and the sensor **170**, the transfer function of the oscillator **190** to be controlled is given by equation (1):

$$G(s) = \frac{A}{(s+s_o)^2 + w_0^2} \cdot e^{T_S \cdot s} = G_0(s) \cdot e^{-T_S \cdot s} \tag{1}$$

In what follows it is assumed that the damping s_0 of the oscillator **190** is much smaller than its resonance angular frequency ($s_0 \ll \omega_0$), and that the oscillator **190** is excited altogether with the harmonic force signal F_e, which has a force amplitude superposing, respectively amplitude modulating an exciting oscillation with the resonance angular frequency ω_0 of the oscillator:

$$F_e = F \cdot \cos(\omega_0 \cdot T) \tag{2}$$

According to FIG. **1**B the actuator **180**, the oscillator **190** and the sensor **170** of FIG. **1**A can then be illustrated as a system with a summation point **191** and a transfer function G(s). A controller signal generated by the controller unit **120** is added to the harmonic force signal F_e at the summation point **191** and the transfer function G(s) acts on the sum signal according to equation (1).

The low pass filter **124** which has to show sufficient damping at the double resonance angular frequency of the oscillator to damp the frequency conversion product sufficiently at $2\cdot\omega_0$, limits the bandwidth of the controller and hence its reaction rate with respect to changes in the force amplitude F.

FIG. **1**C schematically illustrates the signal u(t) at the output of a continuous PI-controller with transfer function $G_R(s)$. A constant input signal $x_d(t)$ at the controller input generates a time proportional gradient u(t) at the controller output.

For a continuous PI-controller with amplification factor K_P and the integral action coefficient K_I the step response u(t) is produced by a step signal $\sigma(t)$ as input signal according to equation (3):

$$u(t) = (K_P + K_I \cdot t) \cdot \sigma(t). \tag{3}$$

By L-transformation of $\sigma(t)$ and equation (3), the transfer function $G_R(s)$ follows:

$$G_R(s) = \frac{U(s)}{X_d(s)} = s \cdot \left(K_P \cdot \frac{1}{s} + K_I \cdot \frac{1}{s^2}\right) = K_P \cdot \frac{s + \frac{K_I}{K_P}}{s} \tag{4}$$

A pole at s=0 resulting from the integral component is characteristic for the continuous PI-controller. In a PI-controller used in connection with a controlled system of first order with a system function $G_s(s)$, the system parameter K_s, and the boundary angular frequency ω_1 is, according to equation (5),

$$G_S(s) = K_S \cdot \frac{1}{s + \omega_1} \tag{5}$$

then the controller parameter amplification factor K_P and integral action coefficient K_I are typically chosen so that the pole in the system function $G_S(s)$ (system pole) is compensated by the zero of the transfer function of the controller $G_R(s)$ (controller zero). Equating coefficients in the equations (4) and (5) results in a condition for the controller parameter given by the relation according to equation (6):

$$\omega_1 = \frac{K_I}{K_P} \tag{6}$$

Equation (6) determines only the ratio of the amplification factor K_P to the integral action coefficient K_I. The product of the system transfer function $G_S(s)$ and controller transfer function $G_R(s)$ gives the transfer function of the corrected open loop $G_k(s)$. As the system pole according to equation (5) and the controller zero according to equation (4) cancel, the transfer function of the corrected open loop $G_k(s)$ is given by the relation according to equation (7).

$$G_k(s) = G_S(s) \cdot G_R(s) = K_S \cdot K_P \cdot \frac{1}{s} \tag{7}$$

From the corrected open loop frequency response, the stability properties of the closed loop can be deduced via the Nyquist criterion. Because of the integral characteristics of the corrected open loop an absolute value characteristic results which declines with 20 db/decade. The phase always amounts to −90° for positive frequencies, to which application of the Nyquist criterion is typically limited. The phase characteristic is an odd function and has, at frequency 0, a 180° step from +90° for negative frequencies to −90° for positive frequencies. The transfer function $G_w(s)$ for the closed loop generally results from that of the corrected open loop $G_k(s)$ according to equation (8):

$$G_w(s) = \frac{G_k(s)}{1 + G_k(s)} \tag{8}$$

From equation (8) it follows that the transfer function $G_w(s)$ for the closed loop is only stable when the locus of the corrected open loop neither encloses nor runs through the point −1 for $0 \le \omega < \infty$. One condition equivalent to this is that, at the transition of the absolute value characteristic of the corrected open loop through the 0 dB line, the phase of the corrected open loop is larger than −180°. As the phase is constant at −90° in the above case, the closed loop is thus always stable independent of the choice of amplification factor K_P.

The bandwidth of the closed loop can be deduced from the frequency at the transition of the absolute value characteristic through the 0 dB line. The absolute value frequency response can be shifted via the amplification factor K_P along the ordinate and, thus, the transition through the 0 dB line, respectively influencing the bandwidth that results from it.

FIG. **1**D illustrates, for one example with a controlled system of first order with the boundary angular frequency $\omega_1 = 2 \cdot \pi \cdot 100$ Hz, a system parameter $K_S = \omega_1$ and a PI-controller whose controller zero is chosen to compensate the system pole and whose amplification factor is $K_P = 2$. The figure shows the absolute value frequency responses of controlled system, controller, corrected open loop, and closed loop in the left column from top to bottom and, in the right column from top to bottom, the phase frequency responses of the controlled system, controller, corrected open loop, and closed loop. As can be seen from the diagram at the bottom left, the bandwidth of the open loop defined by the frequency at which the absolute value frequency response of the closed loop has dropped by 3 dB amounts to approximately 100 Hz.

The use of a classical PI-controller assumes a comparatively constant common variable. For this reason applications in which a harmonic common variable of almost constant frequency is to be controlled require a demodulator and a downstream low pass filter, which generate a corresponding baseband signal from the harmonic input signal.

SUMMARY AND OBJECTS OF THE INVENTION

It is therefore the object of the invention to provide an improved controller concept for resetting the deflection of oscillators of the type that oscillate harmonically in the stationary case (e.g. the deflection of one of the movably supported units of a rotation rate sensor) affected by a disturbance.

The present invention addresses the preceding and other objects by providing, in a first aspect, a controller unit. Such controller unit comprises a PI-controller with a proportional transfer element and an integrating transfer element arranged parallel to the proportional transfer element. A controller input of the controller unit is connected with both transfer elements.

A transfer function of the PI-controller has a conjugate complex pole at a controller angular frequency ω_r in the s-plane or a pole at $e^{\pm j\omega_r T}$ in the z-plane, where T is the sampling time of a discrete input signal of the PI-controller and ω_r is larger than 0.

In a second aspect, the invention provides a device. Such device includes a movably supported oscillator that is excitable to an oscillation with the resonance angular frequency ω_0 along a direction of excitation and a controller unit comprising a PI-controller with a proportional transfer element and an integrating transfer element arranged parallel to the proportional transfer element where a controller input of the controller unit is connected with both transfer elements.

An integral action coefficient of the integrating transfer element and an amplification factor of the proportional transfer element are chosen so that the PI-controller is suitable for generating, at admission with a harmonic input signal of the controller angular frequency ω_r modulated by the step function at the controller input, a harmonic oscillation of the controller angular frequency ω_r with rising amplitude at the controller output. The controller angular frequency ω_r is equal to the resonance angular frequency ω_0.

In a third aspect, the invention provides a rotation rate sensor. The sensor includes a movably supported oscillator that is excitable in a direction of excitation to an oscillation with resonance angular frequency ω_0, and a controller unit comprising a PI-controller with a proportional transfer element and an integrating transfer element arranged parallel to the proportional transfer element controller input of the controller unit is connected with both transfer elements.

The transfer function of the PI-controller has a conjugate complex pole at the resonance angular frequency ω_0 in the s-plane or at $e^{\pm j\omega_0 T}$ in the z-plane, where T is the sampling time of a discrete input signal of the PI-controller (**225**, **325**) and ω_0 is larger than 0.

In a fourth aspect, the invention provides a method for operating a rotation rate sensor. Such method includes the steps of generating a measurement signal by a sensor reproducing a deflection of an oscillator and generating a controller signal for an actuator from the measurement signal, wherein the actuator counteracts the deviation of the oscillator from a harmonic oscillation with the resonance angular frequency ω_0. The controller signal is derived by means of a controller unit from the measurement signal. The controller unit comprises a PI-controller with a proportional transfer element and an integrating transfer element, arranged parallel to the proportional transfer element. A controller input of the controller unit is connected with both transfer elements.

A transfer function of the PI-controller has a conjugate complex pole at the resonance angular frequency ω_0 in the s-plane or a pole at $e^{\pm j\omega_0 T}$ in the z-plane, where T is the sampling time of a discrete input signal of the PI-controller.

In a fifth aspect, the invention provides a method for manufacturing a rotation rate sensor. The method consists of dimensioning a controller unit comprising a PI-controller with a proportional transfer element and an integrating transfer element arranged parallel to the proportional transfer element where a controller input of the controller unit is connected with both transfer elements. The PI-controller is provided with a transfer function that has a conjugate complex pole at a controller angular frequency ω_r in the s-plane or a pole at $e^{\pm j\omega_r T}$ in the z-plane, where T is the sampling time of a discrete input signal of the PI-controller and ω_r is larger than 0. An integral action coefficient of the integrating transfer element and an amplification factor of the proportional transfer element are chosen so that the PI-controller is suitable for generating at admission with a harmonic input signal of controller angular frequency ω_r modulated by the step function at the controller input. A harmonic oscillation of the controller angular frequency ω_r with rising amplitude at the controller output. The controller angular frequency ω_r is chosen so that the controller angular frequency ω_r is equal to the resonance angular frequency ω_0 of an excitation unit of the rotation rate sensor.

The preceding and other features of the invention will become further apparent from the detailed description that follows. Such written description is accompanied by a set of drawing figures in which numerals, corresponding to numerals of the written description, point to the features of the invention. Like numerals point to like features of the invention throughout both the written description and the drawing figures.

BRIEF DESCRIPTION OF THE DRAWINGS

FIG. **1**A is a schematic block diagram of a device with a harmonically excited oscillator and a controller unit for resetting the deflection of the oscillator in accordance with the prior art;

FIG. **1**B is a schematic model of the device of FIG. **1**A;

FIG. **1**C illustrates the transfer function of a continuous PI-controller operated at baseband;

FIG. **1**D comprises multiple frequency response diagrams for a continuous PI-controller for illustration of the operation of controller units in accordance with the prior art;

FIG. **2**A is a schematic block diagram of a device with a controller unit according to one embodiment of the invention that refers to a controller unit with a continuous PI-controller for harmonic command variables and a dead time element;

FIG. **2**B is a schematic illustration of the transfer function of the PI-controller of FIG. **2**A;

FIG. **2**C comprises multiple diagrams for illustrating frequency response for the controller unit illustrated in FIG. **2**A;

FIG. **3**A is a schematic block diagram of a device with a controller unit according to another embodiment of the invention that refers to a controller unit with a discrete PI-controller for harmonic command variables and a dead time element;

FIG. **3**B schematically illustrates the transfer function of the controller unit of FIG. **3**A;

FIG. **3**C comprises multiple frequency response graphs for illustrating the operation of the controller unit illustrated in FIG. **3**A;

FIG. **3**D illustrates a simplified block diagram of the controller unit of FIG. **3**A transformed in the baseband according to one embodiment;

FIG. **3**E comprises a set of schematic frequency response diagrams of a discrete controller unit dimensioned according to one embodiment of a method for manufacturing a controller unit including determination of the controller parameter by eigenvalue specification;

FIG. **4**A is a schematic block diagram of a device with a controller unit according to a embodiment of the invention which refers to a controller unit with a discrete PI-controller for command variables and a controller extension working similarly to a bandpass;

FIG. **4**B schematically illustrates the transfer function of the controller extension according to FIG. **4**A;

FIG. **4**C comprises frequency response diagrams for explanation of the operation of the controller extension according to FIG. **4**A;

FIG. **5**A is a top plan view of the micromechanical portion of a rotation rate sensor according to a further embodiment of the invention;

FIG. **5**B is a cross-sectional view in elevation of the micromechanical portion of the rotation rate sensor of FIG. **5**A;

FIG. **5**C is a schematic block diagram of the rotation rate sensor of FIGS. **5**A and **5**B;

FIG. **6** is a top plan view of the micromechanical part of a rotation rate sensor according to yet another embodiment of the invention;

FIG. **7**A is a simplified flow diagram of a method of operation of a rotation rate sensor in accordance with the invention; and

FIG. **7**B is a flow diagram of a method for manufacturing a rotation rate sensor in accordance with the invention.

DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENT

Turning to the drawings, FIG. **2**A illustrates a device **200** with a control unit **220** which includes a PI-controller **225** for harmonic command variables with an integrating transfer element **222** with an integral action coefficient K_i and a proportional transfer element **224** with an amplification factor K_p. The PI-controller **225** for harmonic command variables generates a harmonic oscillation with the same frequency and time proportional amplitude at the controller output from a harmonic oscillation of constant amplitude at the controller input, which is amplitude modulated with the step function. FIG. **2**B illustrates the transformation of a sine wave modulated step function signal $x_d(t)$ into a harmonic output signal u(t) with time proportional amplitude by the transfer function $G_{R0}(s)$ of the PI-controller **225**. The described behavior of the PI-controller requires dimensioning of the controller parameters K_i, K_p as described below.

Analogous to equation (3), equation (9) below describes the relation between the controller output signal u(t) and the controller input signal $x_d(t)$ for $x_d(t)=\sigma(t)$:

$$u(t)=(K_P+K_I\cdot t)\cdot\sin(\omega_0\cdot t)\cdot\sigma(t). \tag{9}$$

The Laplace-transform of the controller output signal u(t) and controller input signal $x_d(t)$ are derived as in equations (9a) and (9b) below:

$$X_d(s) = \frac{\omega_0}{s^2 + \omega_0^2} \tag{9a}$$

$$U(s) = K_P \cdot \frac{\omega_0}{s^2 + \omega_0^2} + K_I \cdot \frac{2 \cdot \omega_0 \cdot s}{(s^2 + \omega_0^2)^2}. \tag{9b}$$

The transfer function $G_{R0}(s)$ of the PI-controller **225** for harmonic command variables follows as in equation (10) below:

$$G_{R0}(s) = \frac{U(s)}{X_d(s)} = K_P \cdot \frac{s^2 + 2 \cdot \frac{K_I}{K_P} \cdot s + \omega_0^2}{s^2 + \omega_0^2} \tag{10}$$

A conjugate complex pole at $s=\pm j\omega_0$ resulting from the generalized integral component is characteristic of the continuous PI-controller **225**. With an harmonic oscillation of frequency ω_0 at the controller input, the PI-controller **225** generates no phase shift at the controller output. For adjusting an arbitrary phase, the controller unit **220** therefore additionally includes a dead time element **226** with controller dead time T_R in series with the PI-controller **225**. The transfer function $G_R(s)$ of the controller unit **220** thus follows from equation (11) below:

$$G_R(s) = G_{R0}(s) \cdot e^{-T_R \cdot s} = K_P \frac{s^2 + 2 \cdot \frac{K_I}{K_P} \cdot s + \omega_0^2}{s^2 + \omega_0^2} \cdot e^{-T_R \cdot s} \tag{11}$$

The controller parameters K_i, K_P are chosen so that the controller zeros of the controller transfer functions according to equation (11) compensate the conjugate complex system pole in the system transfer functions according to equation (1). Equations (12a) and (12b) result from equating the coefficients of equations (1) and (11) for determination of the controller parameters K_i, K_p:

$$2 \cdot \frac{K_I}{K_P} \stackrel{!}{=} 2 \cdot s_0 \tag{12a}$$

$$\omega_0^2 \stackrel{!}{=} \omega_0^2 + s_0^2. \tag{12b}$$

According to one embodiment, the damping s_0 and the resonance angular frequency ω_0 of the oscillator **190** are chosen so that $s_0 \ll \omega_0$ is satisfied and that, hence, equation (12b) is satisfied in very good approximation. Equation (12c) results from equation (12a) as a dimensioning rule for the ratio of the integral action coefficient K_I to the amplification factor K_P:

$$\frac{K_I}{K_P} \stackrel{!}{=} s_0. \tag{12c}$$

The transfer function $G_k(s)$ of the corrected open loop results from the product of the system transfer function $G_S(s)$ and the controller transfer function $G_R(s)$. As the expression for the conjugate complex system pole and the conjugate complex controller zeros cancel by appropriate dimensioning according to equations (12b), (12c), the transfer function $G_k(s)$ of the corrected open loop results as equation (13):

$$G_k(s) = G_S(s) \cdot G_R(s) = A \cdot K_P \cdot \frac{1}{s^2 + \omega_0^2} \cdot e^{-(T_S + T_R) \cdot s} \tag{13}$$

By feedback control with a conventional PI-controller, a jump from +90° to −90° occurs in the phase frequency response of the corrected open loop at the frequency $\omega=0$. In contrast, a 180° phase jump occurs at the frequency ω_0 in the PI-controller **225** designed for harmonic command variables (not necessarily between +90° and −90°). According to one embodiment, the controller dead time T_R is therefore chosen so that the 180° phase jump occurs as close as possible to ω_0, for example by dimensioning the controller parameters according to equation (14a) below:

$$(T_S + T_R) \cdot \omega_0 = \frac{3}{2} \cdot \pi \tag{14a}$$

If the phase shift produced by the system dead time T_S alone at ω_0 is less than 90°, then the phase ratio of 180° can also be generated by an inverting controller. In this case the phases produced by the controller dead time T_R and the system dead time T_S at ω_0, respectively, must merely add to $\pi/2$. The dimensioning rule for the controller dead time T_R is then:

$$(T_S + T_R) \cdot \omega_0 = \frac{\pi}{2}. \tag{14b}$$

The stability properties of the closed loop can be deduced via the Nyquist criterion from the frequency response of the corrected open loop. The corrected open loop consists of the generalized integrator and the combination of system dead time T_S and controller dead time T_R. By appropriate dimensioning of the controller dead time T_R according to equations (14a) or (14b), the phase characteristics at the frequency ω_0 has a 180° jump between +90° for lower frequencies $\omega<\omega_0$ to −90° to higher frequencies $\omega>\omega_0$. The transfer function $G_w(s)$ of the closed loop is related to that of the corrected open loop $G_k(s)$, again, according to equation (8).

$$G_w(s) = \frac{G_k(s)}{1 + G_k(s)}. \tag{15}$$

When the controller dead time T_R is determined according to equation (14a) the closed loop is exactly stable when the locus of the corrected open loop neither encloses nor runs through the point −1 for $0\leq\omega<\omega_0$. In contrast, when the controller dead time T_R is determined according to equation (14b) and the PI-controller **225** generates a 180° phase then the closed loop is exactly stable when the locus of the corrected open loop at a negative real axis starts at a value larger than −1.

In the interval $0\leq\omega<\omega_0$ the absolute value characteristic intersects the 0 dB line at the gain crossover frequency where the frequency difference from ω_0 at the gain crossover frequency determines the bandwidth of the closed loop. The absolute value frequency response and, hence, the gain crossover frequency can be shifted by the amplification factor K_P along the ordinate so that the resulting bandwidth of the closed loop is adjustable. According to one embodiment, the amplification factor K_P is chosen so that the bandwidth is maximal within the limits given by the stability criteria.

The graphs of the left hand column (from top to bottom) of FIG. 2*c* illustrates the absolute value frequency responses for the controlled system, the controller, the corrected open loop and the closed loop while those of the right column illustrate phase frequency responses for the controlled system, the controller, the corrected open loop and the closed loop for an embodiment having system parameters as follows: resonance angular frequency of the oscillator $\omega_0=2\cdot\pi\cdot9000$ Hz; damping of the oscillator

$$s_0 = \frac{\omega_0}{200000};$$

amplification $A=s_0^2+\omega_0^2$; and system dead time

$$T_s = \frac{\pi}{4\cdot\omega_0}.$$

The controller zero is chosen so that the system pole is compensated. As the phase (at ω_0) which is produced by the system dead time is smaller than 90°, the phase ratio of 180° can be realized by a minus sign in the controller (inverting controller). For an amplification factor $K_P=-1/10$ the integral action coefficient K_I results from equation (12c) and the controller dead time T_R results from equation (14b) as $T_R=\pi/4\cdot\omega_0$.

The resulting bandwidth of the closed loop amounts to approximately 500 Hz and is clearly larger than in the comparative example of a conventional PI-controller operated in the baseband.

The device of FIG. 2A includes an oscillator **190** and a controller unit **220**. The oscillator **190** is a mass movable along a direction of excitation that is capable of oscillating with a resonance angular frequency ω_0. In the stationary case (e.g. without admission with a disturbance) the oscillator **190** undergoes a translational or rotational oscillation with the resonance angular frequency ω_0. A deflection effected by a force amplitude is superposed to this oscillation. A sensor **170** captures the movement of the oscillator **190** and outputs a measurement signal, which reproduces the entire deflection of the oscillator **190** along the direction of excitation. The measurement signal corresponds to a controller input signal for the controller unit **220**. The controller unit **220** generates a controller output signal from the controller input signal and outputs the controller output signal to an actuator **180**. The actuator **180** counteracts the deflection effected by the force signal F of the oscillator **190** and compensates the deflection such that the oscillator **190** performs a harmonic oscillation with constant amplitude at the resonance angular frequency ω_0.

The controller unit **220** has a PI-controller **225**, which includes a proportional transfer element **224** with an amplification factor K_P and a integrating transfer element **222** with an integral action coefficient K_I, for harmonic command variables. The integral action coefficient K_I and the amplification factor K_P are chosen so that the zero of the controller transfer function of the PI-controller **225** and the conjugate complex pole of the system transfer function, which describes the oscillator **190**, compensate in the s-plane.

According to one embodiment, the damping s_0 of the oscillator **190** with respect to deflection in the direction of excitation is much smaller than the resonance angular frequency ω_0 of the oscillator **190** and the ratio of the integral action coefficient K_I to the amplification factor K_P in sec^{-1} corresponds approximately to the damping s_0. Moreover, the amplification factor K_P can be chosen so that the resulting bandwidth is as high as possible for respective stability requirements. The integral action coefficient K_I is then chosen in relation to the damping s_0 and the amplification factor K_P according to equation (12c).

According to one embodiment, the system formed from the actuator **180**, the oscillator **190** and the sensor **170** has a dead time T_S and the controller unit **220** has a dead time element **226** with the controller dead time T_R acting serially to the PI-controller **225**. The controller dead time T_R is chosen in relation to the resonance frequency ω_0 of the oscillator **290** and the system dead time T_S is chosen so that the phase frequency response of the corrected open loop at the frequency ω_0 has a phase jump from +90° to −90° towards higher frequencies.

According to a first variant of this embodiment, the PI-controller for harmonic command variables does not flip the sign and the controller dead time T_R is chosen so that the product of the resonance angular frequency ω_0 and the sum of system dead time T_S and controller dead time T_R is $3\pi/2$. According to another variant of this embodiment, the PI-controller **225** inverts the sign, shifts the phase about 180°, and the phase effected by the controller dead time T_R and the system dead time T_S at the resonance angular frequency ω_0 merely adds to $\pi/2$ so that the product of the resonance angular frequency ω_0 and the sum of system dead time T_S and controller dead time T_R is $\pi/2$.

As the controller unit **220** provides no baseband transformation (which requires a low pass filter for damping higher frequency conversion products), the controller **220** can be formed with a considerable broader band. The controller unit **220** reacts faster to disturbances than comparative controllers that provide a baseband transformation.

FIGS. **3**A to **3**E refer to an embodiment in which the controller **220** has a discrete PI-controller **325** for harmonic command variables with a discrete proportional transfer element **324** of amplification factor K_P and a discrete integrating transfer element **322** with integral action coefficient K_I. According to one embodiment an analog measurement signal from the sensor **170**, is sampled by a sampling unit **321** with a sampling time T and transformed into a digital input signal for the discrete PI-controller **325**. According to another embodiment the sensor **170** already outputs a digital measurement signal.

According to an embodiment in which the system including the actuator **180**, the oscillator **190** and the sensor **170** has a system dead time T_S, the controller unit **220** includes a dead time element **326** arranged in series with the discrete PI-controller **325** with a controller dead time T_R. The system dead time T_S as well as the controller dead time T_R are expressed as multiples of the sampling time T according to equations (16a) and (16b) below:

$$T_S=\beta_S\cdot T \text{ and} \tag{16a}$$

$$T_R\beta_D\cdot T. \tag{16b}$$

In this process, the controller dead time T_R is determined so that the corrected open loop has a phase jump at the resonance angular frequency ω_0 from +90° and −90° towards higher frequencies.

According to one embodiment, the ratio of the integral action coefficient K_I to the amplification factor K_P is adjusted so that the controller zero of the controller transfer function compensates the conjugate complex system pole of the system transfer function in the s-plane. According to another embodiment, controller parameters are chosen so that the transfer function of the closed loop of an equivalent baseband system has a double real eigenvalue. The controller unit **220** is, for example, realized as a digital circuit (e.g., as ASIC (application specific integrated circuit), DSP (digital signal processor) or FPGA (Field Programmable Gate Array)).

FIG. **3**B illustrates the z-transfer function $G_{R0}(z)$ of the discrete PI-controller **325** for harmonic command variables according to FIG. **3**A. The transfer function $G_{R0}(z)$ is determined so that the PI-controller **325** generates, from an input signal $x_d(k)$ including a harmonic oscillation modulated with the step function $\sigma(k)$, a harmonic oscillation of the same frequency with a time proportional amplitude as the controller output signal u(k), defined by equation (17) below:

$$u(k)=(K_P+K_I\cdot T\cdot k)\cdot\sin(\omega_0\cdot T\cdot k)\cdot\sigma(k) \tag{17}$$

The input function $X_d(z)$ and the output function U(z) result from z-transformations according to equations (18a) and (18b) below:

$$X_d(z) = \frac{z\cdot\sin(\omega_0\cdot T)}{z^2-2\cdot\cos(\omega_0\cdot T)\cdot z+1} \tag{18a}$$

$$U(z) = K_P\cdot\frac{z\cdot\sin(\omega_0\cdot T)}{z^2-2\cdot\cos(\omega_0\cdot T)\cdot z+1} + K_I\cdot\frac{T\cdot z^3\cdot\sin(\omega_0\cdot T)-T\cdot z\cdot\sin(\omega_0\cdot T)}{(z^2-2\cdot\cos(\omega_0\cdot T)\cdot z+1)^2} \tag{18b}$$

The transfer function $G_{R0}(z)$ of the discrete PI-controller **325** for harmonic command variables is then, according to equation (18c) below:

$$G_{R0}(z) = \frac{U(z)}{X_d(z)} = \frac{(K_P+K_I\cdot T)\cdot z^2-2\cdot K_P\cdot\cos(\omega_0\cdot T)\cdot z+K_P-K_I\cdot T}{z^2-2\cdot\cos(\omega_0\cdot T)\cdot z+1} \tag{18c}$$

As the generalized integral portion, such a discrete PI-controller has a pole at $z=e^{\pm j\omega_0\cdot T}$ and generates, with a harmonic oscillation of the frequency ω_0 at the input, no phase shift at the output. To be able, nevertheless, to adjust an arbitrary phase, the controller unit **220** is provided with a dead time element **326** of retardation β_D according to one embodiment. The controller transfer function $G_R(z)$ of the controller unit **220** with the dead time element **326** and the discrete PI-controller **325** is derived as equation (19) below:

$$G_R(z) = G_{R0}(z)\cdot z^{-\beta_D} = \frac{(K_P+K_I\cdot T)\cdot z^2-2\cdot K_P\cdot\cos(\omega_0\cdot T)\cdot z+K_P-K_I\cdot T}{z^2-2\cdot\cos(\omega_0\cdot T)\cdot z+1}\cdot z^{-\beta_D} \tag{19}$$

The model of the continuous controlled system according to equation (1) has to be discretized accordingly. To this end in the transfer function G(s) of the controlled system according to equation (1) the system dead time T_S is at first expressed as a multiple of the sampling time T according to equation (16a):

$$G(s) = \frac{A}{(s+s_0)^2+\omega_0^2}\cdot e^{-\beta_S\cdot T_s} = G_0(s)\cdot e^{-\beta_S T_s} \tag{20}$$

Generally a step transfer function G(z) of a discretized model of a continuous controlled system with the transfer function G(s) can be calculated according to equation (21):

$$G(z) = \frac{z-1}{z}\cdot Z\left\{\frac{G(s)}{s}\right\} \tag{21}$$

Employing the following abbreviations according to equations (21a) to (21e)

$$K_S = \frac{A}{s_0^2+\omega_0^2} \tag{21a}$$

$$b_1 = 1-e^{-s_0\cdot T}\cdot\cos(\omega_0\cdot T)-\frac{s_0}{\omega_0}\cdot e^{-s_0\cdot T}\cdot\sin(\omega_0\cdot T) \tag{21b}$$

$$b_2 = e^{-2\cdot s_0\cdot T}-e^{-s_0\cdot T}\cdot\cos(\omega_0\cdot T)+\frac{s_0}{\omega_0}\cdot e^{-s_0\cdot T}\cdot\sin(\omega_0\cdot T) \tag{21c}$$

$$a_1 = 2\cdot e^{-s_0\cdot T}\cdot\cos(\omega_0\cdot T) \tag{21d}$$

$$a_2 = -e^{-2\cdot s_0\cdot T} \tag{21e}$$

the step transfer function G(z) for the oscillator **190** resulting from equations (20) and (21) is, according to equation (22):

$$G(z) = K_s\cdot\frac{b_1 z+b_2}{z^2-a_1\cdot z-a_2}\cdot\frac{1}{z^{\beta_s}} = G_0(z)\cdot\frac{1}{z^{\beta_s}} \tag{22}$$

In one embodiment of the invention, the controller dead time T_R is determined so that the phase frequency response of the compensated open loop has a phase jump from +90° to −90° towards higher frequencies at the resonance angular frequency ω_0. The z-transfer function for the compensated open loop results in analogy to equation (13) from the multiplication of the system transfer function G(z) according to equation (20) with the controller transfer function $G_R(z)$ according to equation (19):

$$G_K(z)=G_0(z)\cdot G_{R0}(z)\cdot z^{-(\beta_S+\beta_D)} \tag{23}$$

Analogous to the equations (14a) and (14b), the controller parameter β_D is chosen such that the transfer function of the corrected open loop $G_k(z)$ has a phase jump from +90° to −90° at the resonance angular frequency ω_0:

$$\left(\beta_S+\beta_D+\frac{1}{2}\right)\cdot\omega_0\cdot T = \frac{3}{2}\cdot\pi \tag{24a}$$

In comparison to equation (14a) one finds an additional part of $\frac{1}{2}\omega_0 T$ with respect to the continuous controller, which represents a retardation that can be traced back to the discretizing of an additional half sampling cycle. As in the case of the continuous controller, a phase jump of 180° can be generated by a minus sign in the controller, provided that the phase shift generated by the system dead time $\beta_S\cdot T$ and the discretization, respectively, are smaller than 90° at the resonance angular frequency ω_0 so that the phases generated by the discretization, the controller dead time $\beta_D\cdot T$ and the system dead time $\beta_S\cdot T$, need only add up to $\pi/2$. Accordingly, the dimensioning rule for β_D results in this case in equation (24b):

$$\left(\beta_S+\beta_D+\frac{1}{2}\right)\cdot\omega_0\cdot T = \frac{\pi}{2}. \tag{24b}$$

The equations (24a) and (24b) normally lead to a non-integral value for β_D. Generally, the controller parameter β_D has an integral part n_D and a remainder $1/a_D$ with $a_D>1$ according to equation (25):

$$\beta_D = n_D+\frac{1}{a_D} \tag{25}$$

According to one embodiment, the integral part n_D can be approximated by a retardation chain in accordance with the length denoted by n_D and the fraction $1/a_D$ of a sampling cycle can be approximated by an all-pass filter of first order according to equation (26):

$$z^{-\frac{1}{a_D}} \approx \frac{\alpha_D \cdot z + 1}{z + \alpha_D}. \tag{26}$$

According to one embodiment, the parameter α_D of the all-pass filter is chosen such that the phase of the exact transfer function $z^{-\alpha_D^{-1}}$ and the phase of the all-pass approximation according to equation (26) coincide at the resonance angular frequency ω_0 as far as possible. From these conditions equation (27) results as a conditional equation for the parameter α_D of the all-pass filter:

$$-\frac{\omega_0 \cdot T}{a_D} = \arctan\left(\frac{\alpha_D \cdot \sin(\omega_0 \cdot T)}{\alpha_D \cdot \cos(\omega_0 \cdot T) + 1}\right) - \arctan\left(\frac{\sin(\omega_0 \cdot T)}{\cos(\omega_0 \cdot T) + \alpha_D}\right) \tag{27}$$

According to one embodiment α_D is determined such that, via nested intervals, the zeros of the function according to equation (28) are determined as follows:

$$f(\alpha_D) = \arctan\left(\frac{\alpha_D \cdot \sin(\omega_0 \cdot T)}{\alpha_D \cdot \cos(\omega_D \cdot T) + 1}\right) - \arctan\left(\frac{\sin(\omega_0 \cdot T)}{\cos(\omega_D \cdot T) + \alpha_D}\right) + \frac{\omega_0 \cdot T}{a_D} \tag{28}$$

The determination of n_D and a_D according to equations (25) and (28) is independent of the method for determining the further controller parameters K_P and K_I.

According to one embodiment of a method for manufacturing a controller unit that includes dimensioning the discrete PI-controller **325** according to FIG. **3**A, the amplification factor K_P and the integral action coefficient K_I of the discrete PI-controller **325** are chosen so that the controller zeros in the controller transfer function $G_R(z)$ according to equation (19) compensate the conjugate complex system pole of the system transfer function G(z) according to equation (22). Equating coefficients of equations (19) and (22) with respect to z^1 leads to the dimensioning rule of equation (29):

$$K_P \overset{!}{=} K_I \cdot T \cdot \frac{e^{-s_0 \cdot T}}{1 - e^{-s_0 \cdot T}} \tag{29}$$

Equating coefficients with respect to z^0 provides the dimensioning rule of equation (30):

$$K_P \overset{!}{=} K_I \cdot T \cdot \frac{1 + e^{-2 \cdot s_0 \cdot T}}{1 - e^{-2 \cdot s_0 \cdot T}}. \tag{30}$$

According to one embodiment, the damping s_0 of the oscillator **190** and the sampling time T are chosen such that $s_0 \cdot T << 1$ holds whereby approximations according to (31a) and (31b) are sufficiently precise:

$$e^{-s_0 \cdot T} \approx 1 - s_0 \cdot T \tag{31a}$$

$$e^{-2 \cdot s_0 \cdot T} \approx 1 - 2 \cdot s_0 \cdot T \tag{31b}$$

Making the approximations according to equations (31a) and (31b), the two independent dimensioning rules according to equations (29) and (30) can be approximated by the dimensioning rule below:

$$K_P \overset{!}{=} K_I \cdot T \cdot \frac{1 - s_0 \cdot T}{s_0 \cdot T}, \text{ respectively } K_I \cdot T \overset{!}{=} K_P \cdot s_0 \cdot T. \tag{32}$$

In one embodiment, the ratio of the integral action coefficient K_I to the amplification factor K_P is set equal (or nearly equal) to the damping s_0 of the oscillator. The dimensioning of the discrete PI-controller **325** according to the described method, which includes the compensation of the system pole by the controller zero, leads to a good reference action of the closed loop.

According to another embodiment of a method for manufacturing a controller unit, which includes the dimensioning of a discrete PI-controller **325**, the integral action coefficient K_I and the amplification factor K_P are determined by suitable eigenvalue specification for a system formed from the discrete PI-controller **325** and a discrete baseband model of the oscillator **190**. For this, a baseband model $G_0'(s)$ equivalent to the oscillation model $G_0(s)$ of equation (1) is at first assumed:

$$G_0'(s) = \frac{A'}{s + s_0} \tag{33}$$

The parameters of the equivalent baseband model according
to equation (33) are determined in accordance with equation (34) so that the absolute value of $G_0'(s)$ at $\omega=0$ coincides with the absolute value of $G_0(s)$ at $\omega=\omega_0$:

$$\left|\frac{A'}{s + s_0}\right|_{\omega=0} \overset{!}{=} \left|\frac{A}{(s + s_0)^2 + \omega_0^2}\right|_{\omega=\omega_0}. \tag{34}$$

According to one embodiment, the oscillator **190** is realized such that $\omega_0 >> s_0$ holds and the relation between the parameters A and A' is closely approximated by equation (35):

$$A' = \frac{A}{2 \cdot \omega_0} \tag{35}$$

For the discretization of equivalent baseband model $G_0'(s)$, equation (36) results, by analogy to equation (21)

$$G_0'(z) = \frac{z - 1}{z} \cdot Z\left\{\frac{G_0'(s)}{s}\right\} \tag{36}$$

From equations (33) and (36) the equivalent discretized baseband model is derived as follows.

$$G_0'(z) = \frac{A'}{s_0} \cdot \frac{1 - e^{-s_0 \cdot T}}{z - e^{-s_0 \cdot T}} \tag{37}$$

FIG. **3**D illustrates the discretized baseband model **190***a* of the oscillator **190** according to equation (37) as well as a controller model **325***a* of the discrete PI-controller **325** for harmonic command variables according to FIG. **3**A with functional subunits. The output of the controller model **325***a* is fed back to the input of the discretized baseband model **190***a*. The functional subunits of the controller model **325***a* can be completely or partially realized by hardware (e.g. by integrated circuits, FPGAs (field programmable gate arrays), ASICs (application specific integrated circuits) or DSPs (digital signal processors)), exclusively realized by software, (implemented, for example, in a DSP or a computer), or realized by a combination of hardware and software.

The system with the discretized baseband model **190***a* and the controller model **325***a* illustrated in FIG. **3**D can be described by a state model according to equations (38a), (38b):

$$\underline{x}(k+1) = \begin{bmatrix} e^{-s_0 \cdot T} - (r_1 + r_2) \cdot \frac{A'}{s_0} \cdot (1 - e^{-s_0 \cdot T}) & -r_2 \cdot \frac{A'}{s_0} \cdot (1 - e^{-s_0 \cdot T}) \\ 1 & 1 \end{bmatrix} \cdot \underline{x}(k+1) + \begin{bmatrix} \frac{A'}{s_0} \cdot (1 - e^{-s_0 \cdot T}) \\ 0 \end{bmatrix} \cdot w(k) \tag{38a}$$

$$\underline{x}(k+1) = \underline{\phi} \cdot \cdot \underline{x}(k) + \underline{h} \cdot w(k) \tag{38b}$$

Calculation of the determinant det(z·I−Φ) leads to the characteristic polynomial of this system according to equation (39b) below:

$$\det(z \cdot \underline{I} - \underline{\Phi}) = \begin{vmatrix} z - e^{-s_0 \cdot T} + (r_1 + r_2) \cdot \frac{A'}{s_0} \cdot (1 - e^{-s_0 \cdot T}) & r_2 \cdot \frac{A'}{s_0} \cdot (1 - e^{-s_0 \cdot T}) \\ -1 & z - 1 \end{vmatrix} \tag{39a}$$

$$\det(z \cdot \underline{I} - \underline{\Phi}) = z^2 - \left(1 + e^{-s_0 \cdot T} - (r_1 + r_2) \cdot \frac{A'}{s_0} \cdot (1 - e^{-s_0 \cdot T})\right) \cdot z + e^{-s_0 \cdot T} - (r_1 + r_2) \cdot \frac{A'}{s_0} \cdot (1 - e^{-s_0 \cdot T}) + r_2 \cdot \frac{A'}{s_0} \cdot (1 - e^{-s_0 \cdot T}) \tag{39b}$$

Calculation of the zeros of the characteristic polynomial according to equation (39b) gives the eigenvalues λ_1, λ_2 of the controlled system, for which the characteristic polynomial can be generally described in the form of equation (40):

$$(z-\lambda_1) \cdot (z-\lambda_2) = z^2 - (\lambda_1+\lambda_2) \cdot z + \lambda_1 \cdot \lambda_2 \tag{40}$$

By equating coefficients between equations (39b) and (40), the controller coefficients depending on the eigenvalues λ_1 and λ_2 (which are to be predetermined) result from equations (41a) and (41b) below.

$$1 + e^{-s_0 \cdot T} - (r_1 + r_2) \cdot \frac{A'}{s_0} \cdot (1 - e^{-s_0 \cdot T}) \overset{!}{=} \lambda_1 + \lambda_2 \tag{41a}$$

$$e^{-s_0 \cdot T} - (r_1 + r_2) \cdot \frac{A'}{s_0} \cdot (1 - e^{-s_0 \cdot T}) + r_2 \cdot \frac{A'}{s_0} \cdot (1 - e^{s_0 \cdot T}) \overset{!}{=} \lambda_1 \cdot \lambda_2 \tag{41b}$$

The equations (41a) and (41b) lead to the equations (42a) and (42b) from which the controller coefficients r_1 and r_2 of the controller model **325***a* can be determined from the parameters of the equivalent discrete baseband model and the predetermined eigenvalues:

$$r_1 = \frac{e^{-s_0 \cdot T} - \lambda_1 \cdot \lambda_2}{\frac{A'}{s_0} \cdot (1 - e^{s_0 \cdot T})} \tag{42a}$$

$$r_2 = \frac{(1 - \lambda_1) \cdot (1 - \lambda_2)}{\frac{A'}{s_0} \cdot (1 - e^{-s_0 \cdot T})} \tag{42b}$$

The amplification factor K_P and the integral action coefficient K_I of the controller unit **220** according to FIG. **3**A are determined from the controller coefficients r_1 and r_2 of the controller model **325***a* of FIG. **3**D according to equations (43a) and (43b):

$$K_P = r_1 \tag{43a}$$

$$K_I T = r_2 \tag{43b}$$

According to one embodiment, the eigenvalues λ_1, λ_2 are predetermined without high dynamics requirements so that the transient oscillation process of the baseband system describes, in close approximation, the envelope of the transient oscillation process of the equivalent bandpass system. In this process, the transferability of the baseband design to the bandpass band only holds approximately, due to the controller dead time of the phase adjustment for the bandpass band system acting as an additional dead time with respect to the baseband system that is not taken into account in the controller design. For this reason, when presetting the eigenvalues with excessively high dynamics requirements, the bandpass band system can be unstable, although the equivalent baseband system is stable. However, by referring to the Nyquist stability criterion, the stability of the bandpass band design can be estimated at any time for the predetermined eigenvalues.

When the method for dimensioning a controller provides presetting of eigenvalues, the position of the two eigenvalues with respect to each other is also predetermined. In contrast, a strong deviation of the two eigenvalues from one other can happen at the dimensioning of the PI-controller for harmonic command variables by pole/zero compensation so that the cancelled system pole remains as eigenvalue in the closed loop and leads to a large time constant at a typically low damping of the oscillator. Indeed the "cancelled" eigenvalue has no influence on the response, but it can be excited by perturbations and can result in long, persistent fading processes. In contrast, the design via eigenvalue presetting allows the presetting of both eigenvalues at approximately the same order of magnitude and thus a positive influence of the perturbation behavior. According to one embodiment, the two eigenvalues are set equal or approximately equal with a maximum deviation of 10% to the larger eigenvalue.

The following exemplary embodiment illustrates the design methods described above for the PI-controller **325** for a controlled system with the following parameters:

$$\omega_0 = 2 \cdot \pi \cdot 9000\,\mathrm{Hz}$$

$$s_0 = \frac{\omega_0}{200000}$$

$$A = s_0^2 + \omega_0^2$$

$$T_S = \frac{\pi}{4 \cdot \omega_0} \quad \beta_S = \frac{\pi}{4 \cdot \omega_0 \cdot T}$$

$$T = \frac{1}{100000\,\mathrm{Hz}}$$

As the phase generated by the system dead time at the resonance frequency ω_0 is less than 90°, a phase ratio of 180° can be realized by an inverting controller (minus sign in the controller). For the controller dead time $\beta_D T$ the dimensioning rule according to equation (44) then results from equation (24b):

$$\beta_D = \frac{\pi}{4\cdot\omega_0\cdot T} - \frac{1}{2} = 0.8\overline{88} \tag{44}$$

According to a method that provides dimensioning of the discrete PI-controller **325** by pole/zero compensation the amplification factor K_P can be, for example, set to $K_P=-1/10$ in analogy to the example illustrated in FIG. **2**C. The integral action coefficient K_I results then from equation (32) to $K_I \cdot T=-2.8274\cdot 10^{-7}$.

FIG. **3**C illustrates, in the left column from top to bottom, absolute value frequency responses for the controlled system, the controller, the corrected open loop, and the closed loop and, in the right column, the corresponding phase frequency responses for the calculated exemplary embodiment. From the frequency responses of the closed loop a bandwidth of about 500 Hz within the 3 dB limits can be read off.

When, in contrast, the discrete PI-controller **325** is dimensioned via eigenvalue presetting, the eigenvalues are chosen for example equally large and according to absolute value, such that the closed loop of the equivalent baseband system has a double real eigenvalue at $\lambda_1=\lambda_2=0.98$. The controller coefficients $r_1=0.14004655$ and $r_2=1{,}41471261\cdot 10^{-3}$ result from equations (42a) and (42b). Taking into account the minus sign required for the phase adjustment, the values for the amplification factor K_P and the integral action coefficient K_I of the discrete PI-controller **325** are $K_P=-0.14004655$ and $K_I \cdot T=-1{,}41471261\cdot 10^{-3}$.

FIG. **3**E illustrates in the upper diagram the step response of the bandbass band system controlled via such a discrete PI-controller as continuous line as well as the step response of the equivalent baseband system as dotted line, which corresponds approximately to the upper branch of the envelope of the step response of the discrete PI-controller. Lower left the absolute value frequency response of the closed bandpass band system and right next to it the corresponding phase frequency response is illustrated from which for example the bandwidth of the closed loop can read off.

FIGS. **4**A to **4**C refer to an embodiment at which the controller unit **320** has a controller extension **328**, arranged in series with the PI-controller **325**, and a dead time element **326** according to FIG. **3**A. In the following, the structure of the controller extension **328** is deduced from an analog controller extension for the baseband.

The oscillator **190** can have further resonances beside the resonance angular frequency at ω_0, such as mechanic structure resonances above or below the resonance angular frequency ω_0. The controller extension **328** is formed such that these further resonances are more strongly damped. To this end, a retardation element of first order (PT_1-element) with a further pole at the kink frequency beyond the desired bandwidth would be added to a conventional PI-controller in the baseband. This additional controller pole causes the controller to no longer act as a proportional element for high frequencies, but its absolute value frequency drops down with 20 db/decade. The step response y(k) of such an extension in the baseband results from the step function σ(k) as input signal u(k) according to equation (45):

$$y(k) = \left(1 - e^{-\frac{k\cdot T}{T_1}}\right)\cdot\sigma(k) \tag{45}$$

The z transform U(z) of the input signal u(k) corresponds to the z transform of the step signal:

$$U(z) = \frac{z}{z-1} \tag{46a}$$

The z transform Y(z) of the output signal y(k) then results:

$$Y(z) = \frac{z}{z-1} - \frac{z}{z - e^{-\frac{T}{T_1}}} \tag{46b}$$

The transfer function $G_{RE0}(z)$ of such a controller extension in the baseband is derived, thus, in analogy to equation (10):

$$G_{RE0}(z) = \frac{1 - e^{-\frac{T}{T_1}}}{z - e^{-\frac{T}{T_1}}} \tag{47}$$

According to one embodiment, the controller extension **238** in the bandpass band is configured now in analogy to the controller extension in the baseband such that the controller extension **328** responses to an admission with a harmonic oscillation of a resonance angular frequency ω_0 modulated by the step function with a harmonic oscillation of the same frequency, wherein the step response of the baseband extension defines the envelope as illustrated on the right side of FIG. **4**B.

FIG. **4**B illustrates the transformation of a sign modulated step function u(k) onto an output signal with sign oscillation whose envelope results from the step response according to the transfer function $G_{RE0}(z)$ of the discrete controller extension in the bandpass band. The input signal of the controller extension **328** in the bandpass band with the transfer function $G_{RE}(z)$ is expressed in equation (48):

$$u(k)=\sin(\omega_0\cdot T\cdot k)\cdot\sigma(k) \tag{48}$$

The controller output signal y(k) is a harmonic oscillation whose envelope corresponds to the step response of the PT_1-controller extension in the baseband:

$$y(k) = \left(1 - e^{-\frac{k\cdot T}{T_1}}\right)\cdot\sin(\omega_0\cdot T\cdot k)\cdot\sigma(k) \tag{49}$$

The z-transform U(z) and Y(z) are set forth in equations (50a) and (50b) below:

$$U(z) = \frac{z\cdot\sin(\omega_0\cdot T)}{z^2 - 2\cdot\cos(\omega_0\cdot T)\cdot z + 1} \tag{50a}$$

$$Y(z) = z\cdot\frac{\sin(\omega_0\cdot T)}{z^2 - 2\cdot\cos(\omega_0\cdot T)\cdot z + 1} - z\cdot\frac{e^{-\frac{T}{T_1}}\cdot\sin(\omega_0\cdot T)}{z^2 - 2\cdot e^{-\frac{T}{T_1}}\cdot\cos(\omega_0\cdot T)\cdot z + e^{-2\frac{T}{T_1}}} \tag{50b}$$

The transfer function $G_{RE}(z)$ of the controller extension **328** for the bandpass band follows:

$$G_{RE}(z) = \frac{Y(z)}{U(z)} = \frac{\left(1 - e^{-\frac{T}{T_1}}\right) \cdot z^2 - e^{-\frac{T}{T_1}} \cdot \left(1 - e^{-\frac{T}{T_1}}\right)}{z^2 - 2 \cdot e^{-\frac{T}{T_1}} \cdot \cos(\omega_0 \cdot T) \cdot z + e^{-2\frac{T}{T_1}}} \tag{51}$$

The controller extension **328** with the transfer function $G_{RE}(z)$ acts in series with the discrete PI-controller **325** similarly to a bandpass of first order with the resonance frequency ω_0 as midband frequency. The absolute value and phase of the compensated open loop at the resonance angular frequency ω_0 in a narrow region around the resonance angular frequency ω_0 according to equation (52) remain unchanged.

$$\omega_0 - \frac{1}{T_1} \leq \omega \leq \omega_0 + \frac{1}{T_1} \tag{52}$$

In this region, the absolute value frequency response of the compensated open loop is barely affected, while, out of this region, a considerable drop of the absolute value happens such that possible undesired resonances can be dropped. FIG. **4**C illustrates the absolute value frequency response as well as the phase frequency response for the controller extension for $T_1=1/(2\cdot\pi\cdot1000\text{ Hz})$ with a transfer function $G_{RE}(z)$ according to equation (51).

FIGS. **5**A to **5**C refer to a micromechanical rotation rate sensor **500** according to a further embodiment. The rotation rate sensor **500** includes an excitation unit **590**, e.g. an excitation frame, suspended at first spring elements **541**. The first spring elements **541** couple the excitation unit **590** to an attachment structure **551** which is fixedly connected to a support substrate **550** illustrated in FIG. **5**B. The spring elements **541** only weakly damp a deflection of the excitation unit **590** with respect to the support substrate **550** along the direction of excitation **501**. A detection unit **580** is coupled to the excitation unit **590** over second spring elements **542** and is movable with respect to the excitation unit **590** mainly along a detection direction **502** orthogonal to the direction of excitation **501**. The direction of excitation **501** and the detection direction **502** run parallel to a surface of the support substrate **550**. The first and second spring elements **541**, **542** are, for example, beam-like structures with small cross sections, which are formed between each of the structures to be coupled.

According to one embodiment, the rotation rate sensor **500** includes first force transmission and sensor units **561**, **571** (e.g. electrostatic force transmitters and sensors) which excite the system formed from the excitation unit **590** and the detection unit **580** to an oscillation along the direction of excitation **501** and/or are able to capture a corresponding deflection of the excitation unit **590**. The rotation rate sensor **500** includes further second force transmission and sensor units **562**, **572** (e.g. electrostatic force transmitters and sensors) which act on the detection unit **580** and/or are able to capture its deflection. According to one embodiment at least one of the second force transmission and sensor units **562**, **572** is controlled such that it counteracts a deflection of the detection unit **580**, caused by a disturbance or, in a closed loop system, caused by a measured variable.

During operation of the rotation rate sensor **500**, the first force transmission and sensor units **561**, **571** excite, for example, the excitation unit **590** to an oscillation along the direction of excitation **501**, wherein the detection unit **580** moves approximately with the same amplitude and phase with the excitation unit **590**. When the arrangement is rotated around the axis orthogonal to the substrate plane, a Coriolis force acts on the excitation unit **590** and the detection unit **580**, which deflects the detection unit **580** with respect to the excitation unit **590** in the detection direction **502**. The second force transmission and sensor units **562**, **572** capture the deflection of the deflection unit **580** and, thus, the rotational movement around the axis orthogonal to the substrate plane.

According to one embodiment, at least one of the force transmission and sensor units **561**, **572**, **562**, **572** acts as actuator and either the excitation unit **590** or the detection unit **580** as oscillator within the meaning of one of the devices **200** described above.

According to one embodiment illustrated in FIG. **5**C of the rotation rate sensor **500**, the first force transmission and sensor units **561**, **571** excite the excitation unit **590** to oscillation with the resonance angular frequency ω_0 along the direction of excitation **501**. In a control loop according to the above discussion, an oscillation of the detection unit **580** along the detection direction **502** (x2-oscillator) can then for example correspond to the harmonic force signal as described above.

The deflection of the x2-oscillator can be captured via the charge on the common movable electrode, which is formed on the excitation unit **590**. The charge can be measured via the attachment structure **551**. A charge amplification unit **521** amplifies the measured signal. While typically a demodulation unit modulates the measured signal with a frequency which corresponds for example to the resonance angular frequency ω_0 before it is fed into a controller unit, the embodiments of the invention feed the non-demodulated harmonic signal as measurement signal within the meaning described above into a controller unit **520** according to the above.

The damping s_0 for the oscillation is considerably smaller than the resonance angular frequency ω_0. The signal measured over the excitation frame the excitation unit **590** partly reproduces the movement of the excitation unit **590** along the direction of excitation **501**. A disturbance whose source can be outside of the rotation rate sensor **500**, or, in a closed loop system, the measurement variable, respectively, superposes the oscillation and modulates its amplitude. The controller unit **520** senses from the modulated harmonic signal a control signal for the second force transmission and sensor units **562**, **572** which counteracts the deflection effected by the disturbance or the measurement variable, respectively. An amplification unit **522** transforms the control signal in a suitable reset signal for the electrodes of the second force transmission and sensor units **562**, **572**. The controller unit **520** is formed and dimensioned according to one of the above described controller units **220**. When the amplitude modulation of the harmonic signal reproduces the measurement variable, a demodulation unit can be provided, which generates the rotation rate signal by demodulation of the harmonic control signal with the resonance angular frequency ω_0.

The rotation rate sensor **505** illustrated in FIG. **6** differs from the rotation rate sensor **500** illustrated in FIG. **5**A by a Coriolis unit **585** arranged between the excitation unit **590** and the detection unit **580**. Second spring elements **542** which couple the Coriolis unit **585** to the excitation unit **590** allow for a deflection of the Coriolis unit **585** relative to the excitation unit **590** in the detection direction **502**. Third spring elements **543**, which can be partly connected with the support substrate **550**, couple the detection unit **580** to the Coriolis unit **585** so that the detection unit **580** can follow the movement of the Coriolis unit **585** along the detection direction **502**, but cannot follow movements along the direction of excitation **501**. The detection unit **580** is fixed with respect to the direction of excitation **501** and is moveable along the detection direction **502**.

According to another embodiment, at least one of the first or second force transmission and sensor units **561**, **562**, **571**, **572** acts as actuator and either the excitation unit **590** or the detection unit **580** or the excitation unit **590** as well as the detection unit **580** act as oscillator according to one of the devices described above, which are operated according to the principle of the bandpass controller. In this process the force transmission and sensor units **561** and **571** act as force transmission and sensor units respectively for the x1-oscillator and the force transmission and sensor units **562** and **572** act as force transmission and sensor units respectively for the x2-oscillator.

A rotation rate sensor according to another embodiment includes two of the arrangements as illustrated in FIG. **5**A or FIG. **6**, which are coupled to each other such that the excitation units perform opposing oscillations in the stationary state with respect to each other. Another embodiment covers rotation rate sensors with four of the arrangements as illustrated in FIG. **5**A or FIG. **6** which are coupled to each other such that every two of the excitation units perform opposing oscillations in the stationary state.

A further embodiment refers to the controller unit **220** illustrated in FIGS. **2**A, **3**A and **4**A. The controller unit **220** includes at least one PI-controller **225**, **325** for harmonic command variables, which has a proportional transfer element **224**, **324** and an integrating transfer element **222**, **322** arranged in parallel to the proportional transfer element **224**, **324**, wherein a controller unit of the controller unit **220** is connected with both transfer elements **222**, **224**, **322**, **324**. The transfer function of the PI-controller **225**, **325** for harmonic command variables has a conjugate complex pole at a controller angular frequency ω_r in the s-plane or at $e^{\pm j\omega_r T}$ in the z-plane, wherein T is the sampling time of a discrete input signal of the PI-controller **325** and wherein ω_r is larger than 0.

To achieve this, the integral action coefficient of the integrating transfer elements **222**, **322** and a amplification factor of the proportional transfer elements **224**, **324** is chosen such that the PI-controller **225**, **325** for harmonic command variables is suitable for generating a harmonic oscillation of the controller angular frequency □$_r$ with rising amplitude at a controller output, with an harmonic input signal of the controller angular frequency ω_r modulated by the step function at the controller input.

The PI-controller **225**, **325** for harmonic command variables can also be employed for a controller derived from a conventional PI-controller for stationary command variables and differs from it by the position of the poles in the s- or z-plane, respectively.

FIG. **7**A refers to a method for operating a rotation rate sensor. During operation of a rotation rate sensor, a sensor generates a measurement signal, which reproduces a deflection of an oscillator (**702**). A controller unit generates a control signal from the measurement signal for an actuator, which counteracts the deviation of the deflection of the oscillator from a harmonic oscillation with the resonance angular frequency ω_0 (**704**). The controller unit has a PI-controller with a proportional transfer element and an integrating transfer element arranged parallel to the proportional transfer element, wherein a controller input of the controller unit is connected with both transfer elements. The transfer function of the PI-controller has a conjugate complex pole at a resonance angular frequency ω_0 of the oscillator in the s-plane or a pole at $e^{\pm j\omega_r T}$ in the z-plane. T is the sampling time of a discrete input signal of the PI-controller and ω_0 is larger than zero.

FIG. **7**B refers to a method for manufacturing a rotation rate sensor. The method includes dimensioning a controller unit with the PI-controller with a proportional transfer element and an integrating transfer element arranged parallel to the proportional transfer element, wherein a controller input of the controller unit is connected with both transfer elements. The PI-controller is provided with a transfer function, which has a conjugate complex pole at a controller angular frequency ω_r in the s-plane or a pole at $e^{\pm j\omega_r T}$ the z-plane, wherein T is the sampling time of a discrete input signal of the PI-controller and ω_r is larger than 0. The controller angular frequency ω_r is chosen such in this process that the controller angular frequency ω_r is equal to the resonance angular frequency ω_0 of an oscillator of the rotation rate sensor (**754**). To this end an integral action coefficient of the integrating transfer element and an amplification factor of the proportional transfer element are chosen such that the PI-controller is suitable for generating a harmonic oscillation of the angular resonance frequency ω_0 with rising amplitude at the controller output, at admission of the harmonic input signal of the resonance angular frequency ω_0 modulated with the step function at the controller input.

While the invention has been described with reference to a presently preferred embodiment, it is not limited thereto. Rather it is limited only insofar as it is defined by the following set of patent claims and includes within its scope all equivalents thereof.

What is claimed is:

1. A controller unit comprising a PI-controller with a proportional transfer element and an integrating transfer element, arranged parallel to the proportional transfer element, wherein a controller input of the controller unit is connected with both transfer elements, characterized in

that a transfer function of the PI-controller has a conjugate complex pole at a controller angular frequency ω_r in the s-plane or a pole at $e^{\pm j\omega_r T}$ in the z-plane, wherein T is the sampling time of a discrete input signal of the PI-controller and ω_r is larger than 0.

2. The controller unit according to claim **1**, wherein

an integral action coefficient of the integrating transfer element and an amplification factor of the proportional transfer element are chosen such that the PI-controller is suitable for generating, at admission with a harmonic input signal of the controller angular frequency ω_r modulated by the step function at the controller input, a harmonic oscillation of the controller angular frequency ω_r with rising amplitude at the controller output.

3. A device comprising a moveably supported oscillator, which is excitable to an oscillation with the resonance angular frequency ω_0 along a direction of excitation, and

the controller unit according to claim **2**, wherein the controller angular frequency ω_r is equal to the resonance angular frequency ω_0.

4. The device according to claim **3**, wherein

the integral action coefficient and the amplification factor are chosen such that the zeros of the transfer function of the PI-controller compensate poles of a transfer function of the oscillator.

5. The device according to claim **3**, wherein

the PI-controller is a continuous PI-controller and

the ratio of integral action coefficient to amplification factor is equal to a damping s_0 of the oscillator in the direction of excitation.

6. The device according to claim **5**, wherein

a controlled system comprising the oscillator has a system dead time T_S,

the controller unit comprises in series to the PI-controller a dead time element with a controller dead time T_R, and either

the PI-controller is an inverting controller and the product of the resonance angular frequency ω_0 and the sum of the system dead time T_S and the controller dead time T_R is equal to $\pi/2$ or

the PI-controller (**225**) is a non-inverting controller and the product of the resonance angular frequency ω_0 and the sum of the system dead time T_S and the controller dead time T_R is equal to $3\pi/2$.

7. The device according to claim **3**, wherein

the PI-controller is a discrete PI-controller, which is admittable with a discrete input signal emerging from a sampling with the sampling time T,

the oscillator has a damping s_0 in the direction of excitation, and the ratio of the integral action coefficient to the amplification factor equals the ratio $s_0:(1-s_0 \cdot T)$.

8. The device according to claim 3, wherein
the PI-controller is a discrete PI-controller, which is admittable with a discrete input signal emerging from a sampling with the sampling time T,
the integral action coefficient and the amplification factor are chosen such that the transfer function of the closed loop of an equivalent baseband system has a double real eigenvalue.

9. The device according to claim 7, wherein
a controlled system comprising the oscillator has a system dead time $\beta_s \cdot T$,
the controller unit comprises a dead time element with a controller dead time $\beta_D \cdot T$ in series to the discrete PI-controller, and either
the discrete PI-controller is an inverting controller and the product from the resonance angular frequency ω_0 and the sum of the system dead time $\beta_s \cdot T$, controller dead time $\beta_D \cdot T$, and the half sampling time T/2 is equal to $\pi/2$ or
the discrete PI-controller is a non-inverting controller and the product of the resonance angular frequency ω_0 and the sum of the system dead time $\beta_s \cdot T$, the controller dead time $\beta_D \cdot T$, and the half sampling time T/2 is equal to $3\pi/2$.

10. The device according to claim 3, characterized by
a controller extension, arranged in series to the PI-controller, which acts as bandpass with a midband frequency at the resonance angular frequency ω_0.

11. The device according to claim 10, wherein
a transfer function $G_{RE}(z)$ of the controller extension with a bandwidth $1/T_1$ is determined by the following equation:

$$G_{RE}(z) = \frac{\left(1 - e^{-\frac{T}{T_1}}\right) \cdot z^2 - e^{-\frac{T}{T_1}} \cdot \left(1 - e^{-\frac{T}{T_1}}\right)}{z^2 - 2 \cdot e^{-\frac{T}{T_1}} \cdot \cos(\omega_0 \cdot T) \cdot z + e^{-2\frac{T}{T_1}}}$$

12. The device according to claim 3, wherein
the device is a rotation rate sensor and the oscillator is an excitation unit, a Coriolis unit or a detection unit of the rotation rate sensor, wherein
the excitation unit is deflectable by a force transmitter along a direction of excitation and is suitable for an oscillation with the resonance angular frequency ω_0,
the Coriolis unit is attached to the excitation unit such that the Coriolis unit follows a movement of the excitation unit along the direction of excitation and that the Coriolis unit is additionally moveable along a detection direction orthogonal to the direction of excitation, and
the detection unit is attached such to the excitation unit or to the Coriolis unit that the detection unit either
follows a movement of the excitation unit along the direction of excitation and is additionally moveable along a detection direction orthogonal to the direction of excitation, or
follows a movement of the Coriolis unit along a detection direction orthogonal to the direction of excitation and is fixed along the direction of excitation.

13. A rotation rate sensor comprising
a moveably supported oscillator which is excitable in a direction of excitation to an oscillation with a resonance angular frequency ω_0, and
a controller unit comprising a PI-controller with a proportional transfer element and an integrating transfer element, arranged parallel to the proportional transfer element, wherein a controller input of the controller unit is connected with both transfer elements, characterized in that
the transfer function of the PI-controller has a conjugate complex pole at the resonance angular frequency ω_0 in the s-plane or at $e^{\pm j\omega_0 T}$ in the z-plane, wherein T is the sampling time of a discrete input signal of the PI-controller and ω_0 is larger than 0.

14. The rotation rate sensor according to claim 13, wherein
the oscillator is an excitation unit which is deflectable by a force transmitter along a direction of excitation and is suitable for an oscillation with the resonance angular frequency ω_0.

15. A method for operating a rotation rate sensor, comprising
generating a measurement signal by a sensor reproducing a deflection of an oscillator, and
generating a controller signal for an actuator from the measurement signal, wherein the actuator counteracts the deviation of the oscillator from a harmonic oscillation with the resonance angular frequency ω_0, the controller signal is deduced by means of a controller unit from the measurement signal, and the controller unit comprises a PI-controller with a proportional transfer element and an integrating transfer element, arranged parallel to the proportional transfer element, wherein a controller input of the controller unit is connected with both transfer elements, wherein
a transfer function of the PI-controller has a conjugate complex pole at the resonance angular frequency ω_0 in the s-plane or a pole at $e^{\pm j\omega_0 T}$ in the z-plane, wherein T is the sampling time of a discrete input signal of the PI-controller.

16. A method for manufacturing a rotation rate sensor, comprising
dimensioning a controller unit comprising a PI-controller with a proportional transfer element and an integrating transfer element, arranged parallel to the proportional transfer element, wherein a controller input of the controller unit is connected with both transfer elements, characterized in
that the PI-controller is provided with a transfer function that has a conjugate complex pole at a controller angular frequency ω_r in the s-plane or a pole at $e^{\pm j\omega_r T}$ in the z-plane, wherein T is the sampling time of a discrete input signal of the PI-controller and ω_r is larger than 0 and an integral action coefficient of the integrating transfer element and an amplification factor of the proportional transfer element are chosen such that the PI-controller is suitable for generating, at admission with a harmonic input signal of controller angular frequency ω_r modulated by the step function at the controller input, a harmonic oscillation of the controller angular frequency ω_r with rising amplitude at the controller output, and
the controller angular frequency ω_r is chosen such that the controller angular frequency ω_r is equal to the resonance angular frequency ω_0 of an excitation unit of the rotation rate sensor.

17. The method according to claim 16, wherein
the integral action coefficient and the amplification factor are chosen such that the zeros of the transfer functions of the PI-controller compensate the poles of the transfer functions of the excitation unit.

18. The method according to claim 16, wherein
the PI-controller is a discrete PI-controller, which is admittable with a discrete input signal emerged from a sampling with sampling time T, and
the integral action coefficient and the amplification factor are chosen such that the transfer function of the closed loop of an equivalent baseband system has a double real eigenvalue.

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